RESEARCH DEPARTMENT

Television sound transmission by modulated pulses in the line-blanking intervals

RESEARCH REPORT No. T-176

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TELEVISION SOUND TRANSMISSION BY MODULATED PULSES IN THE LINE BLANKING INTERVALS

Page 4:

Equation (1) is shown in the wrong position (bottom of first column) and should follow the first paragraph of Section 3.2.1.

Page 28; second column, third paragraph, first line:

For "poses" read "posed".

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SUMMARY

This report reviews the properties of modulated pulses with particular reference to their application to television sound transmission. It is shown that combinations of modulated pulses and video signals could prove practicable for both broadcasting and link purposes.

1. INTRODUCTION

Up to the present, broadcast television systems have utilized sound channels based upon either amplitude or frequency modulation of a continuous carrier, although a number of proposals have been made in the past for incorporating the sound within the video waveform by means of time-division multiplex, using modulated pulses or bursts occurring during otherwise unoccupied intervals (e.g. line-blanking intervals). Proposals of this nature appeared in technical papers and patent specifications as far back as 1933.[1,2]

In 1946 details of a proposed practical system were published[3] which was based upon the use of duration-modulated pulses occurring within the line-synchronizing intervals, the amplitude of the pulses being some 3 dB greater than the video-signal excursion from synchronizing level to white level. The system was tested and demonstrated as applied to the 405-line system, which resulted in a rather unsatisfactory upper audio-frequency limit (i.e. somewhat less than 5 kHz). A further and probably more important drawback associated with this particular proposal was its performance in the presence of fairly high levels of noise. In an attempt to show a marked economic advantage over the conventional a.m. sound system, the sound pulses at the receiver were not extracted from the video waveform by gating prior to demodulation. This necessitated the use of a limiter operating at a level near to that of the peak of the sound pulse and caused the signal-to-noise ratio of the receiver sound output to become very unsatisfactory at a receiver-input signal level that was quite usable in terms of picture signal.

Since the appearance of the above-mentioned paper, little attention has been paid to the application of modulated pulses as a means for television-sound broadcasting.

In recent years, however, the widespread development of television in the United Kingdom has led to a situation in which the space available in the radio-frequency Bands I and III might not be adequate to provide a sufficient number of channels for satisfactory coverage of the country with two programmes if 625-line transmissions were used in conjunction with a conventional a.m. or f.m. sound channel. It has been suggested that this problem could be eased, and perhaps solved, if a satisfactory sound-pulse system were developed. It has also been suggested that, even if such a system were not used for broadcasting, it could enable satisfactory sound signals to be conveyed, say, from a programme-origination point to the main transmitter sites; the sound signals would be deleted from the vision waveform prior to emission. This form of sound-signal distribution could appreciably reduce the cost of sound circuits and could lead to a reliability of service greater than that obtained at present, where breakdown of a separate sound link tends to occur more frequently than breakdown of a vision circuit; systems of this nature have been tested in France[4] and the U.S.S.R.[5]

This report outlines some of the general properties of systems whereby sound signals are conveyed as modulated pulses located in line-blanking intervals and discusses their advantages and disadvantages in terms of the applications mentioned previously.

2. GENERAL

The effect of sampling a single tone by regularly recurring pulses having durations small compared with the recurrence period is well known; the

process may be illustrated by Fig. 1 which shows the lower-frequency part of the spectrum resulting from the sampling process. It will be seen that the spectrum includes a component at ω_a representing the single tone, together with symmetrical pairs of components differing from the pulse recurrence frequency $\omega_r/2\pi$ and its harmonics by the frequency of the tone. It is possible to recover the tone provided that two conditions are satisfied:

(*a*) The frequency of the tone is restricted, before sampling, to a maximum of half the recurrence frequency.

(*b*) The samples are passed either to a low-pass filter cutting off at half the recurrence frequency or to a band-pass filter and detector where the pass-band of the filter is centred on the recurrence frequency, or one of its harmonics, and extends either side of centre frequency by half the recurrence frequency.

Failure to observe either of these two conditions results in the appearance of an "inversion" of the tone having a frequency equal to the difference between the frequency of the tone and the pulse-recurrence frequency.

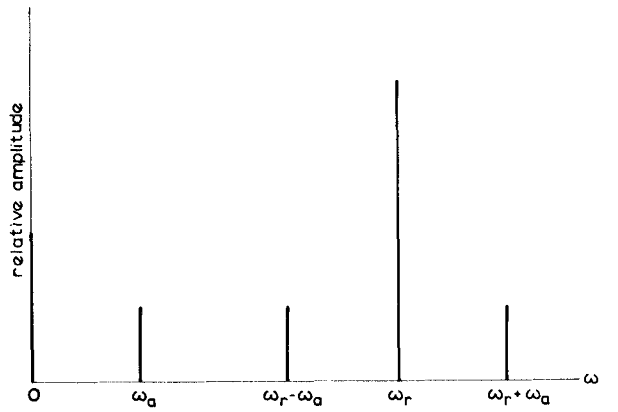

Fig. 1 - Lower frequencies of spectrum produced by sampling an a.f. wave of frequency $\omega_a/2\pi$ by narrow pulses having a recurrence frequency of $\omega_r/2\pi$

Thus, in any combined sound and vision system in which single samples of the sound waveform are transmitted once per line scanning period, the upper audio-frequency limit is equal to half the line-scan frequency; for the 625 line and 525 line standards, the practical upper limit is likely to be rather less than 7·8 kHz. If sampling of the sound waveform occurs at twice the line-scan frequency, and alternate samples are delayed so that pairs of samples are transmitted during each line-blanking interval, the upper audio-frequency limit may be doubled.

Although the characteristics of sampling have been outlined in terms of a single tone, they apply equally well when many tones are present simultaneously (i.e. a complex wave) provided that the above-mentioned conditions (*a*) and (*b*) are satisfied.

Any system of sound transmission involving the use of pulses which, at least nominally, recur regularly, involves a sampling process similar to that described. Some characteristic of each pulse is varied according to the amplitude of a sample of the audio-frequency wave;, usually the sampling action and the generation of the modulated pulses are coincident. Fig. 2 shows one form of pulse, occurring within the line-blanking interval, which could be modulated by a.f. signals.

3. TYPES OF MODULATED PULSES

Many forms of modulated pulse are known. However, attention is confined in this report to the two general forms that are potentially suitable for both the applications mentioned in Section 1. The two forms of pulse modulation are:

(*a*) Pulse-amplitude modulation and

(*b*) Pulse-time modulation, which includes pulse-position modulation and pulse-duration modulation.

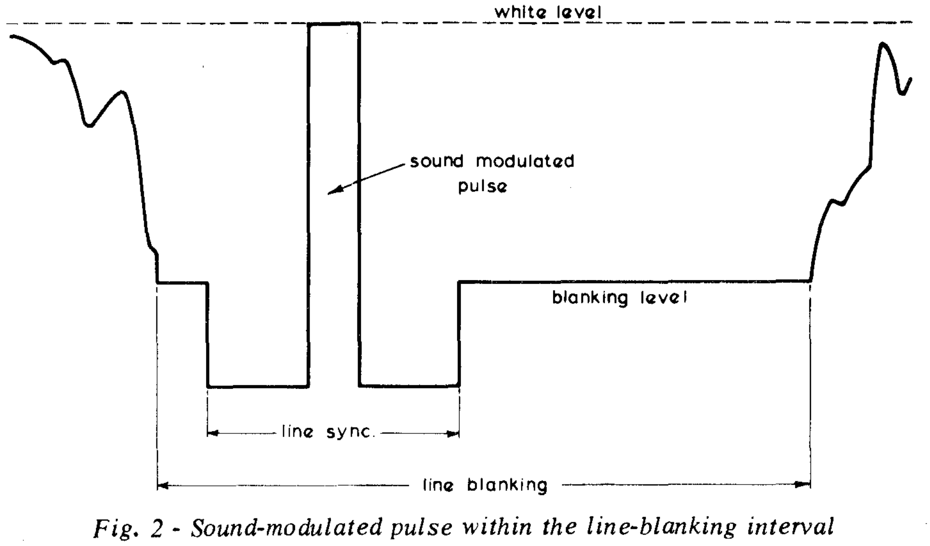

Fig. 2 - Sound-modulated pulse within the line-blanking interval

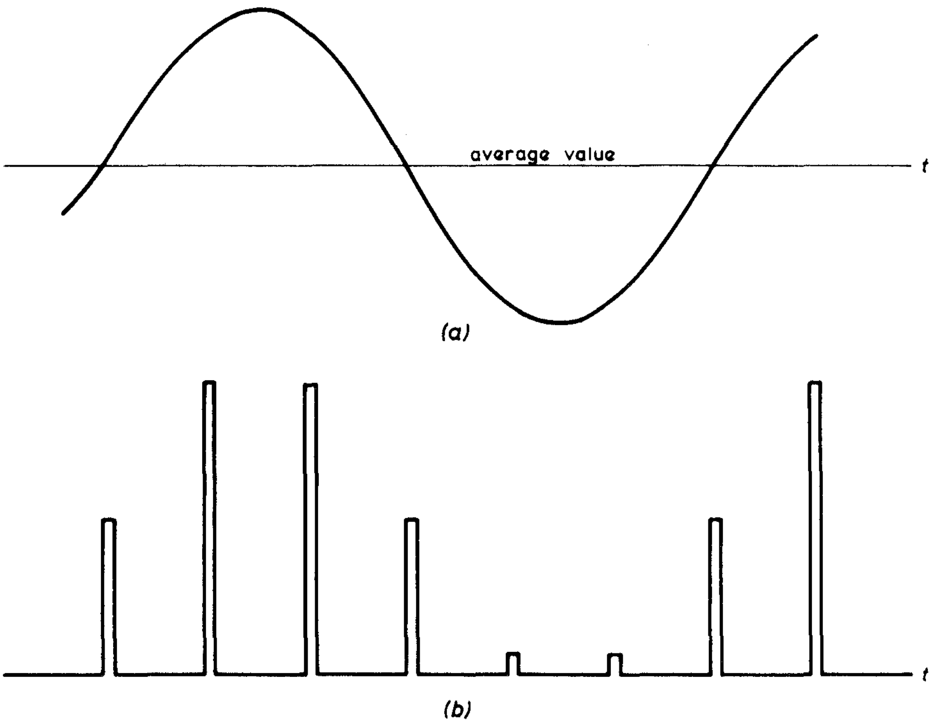

Fig. 3 - P.A.M.
(*a*) Modulating Wave (*b*) Pulses amplitude-modulated by (*a*)

3.1. Pulse-amplitude Modulation (P.A.M.)

In this arrangement, the amplitude of each pulse of a regularly recurring train is linearly related to the magnitude of the a.f. signal at the time of occurrence of the pulse. The pulses may, on the one hand, exactly describe the a.f. wave during each pulse interval (i.e. the waveform of the pulse between its leading and trailing edges follows the variations of the a.f. wave) or, on the other hand, only describe the magnitude of the modulating wave at a certain regularly recurring instant within the duration of each pulse; in the latter case the instant of sampling usually coincides with the leading edge of the pulse. However, no significant difference between the two forms of P.A.M. arises when the duration of each pulse is very short compared with the period of the highest modulating frequency (i.e. half the pulse-recurrence frequency), as is the case for a train of modulated pulses suitable for inclusion in a video waveform. Fig. 3 illustrates a waveform corresponding to pulse-amplitude modulation. It will be seen that a train of amplitude-modulated pulses may be regarded as a train of samples of the a.f. wave; thus its spectrum is identical with that shown in Fig. 1.

Within the audio range set by the recurrence frequency and the design of practicable low-pass filters, the system does not introduce any intrinsic distortion.

3.2. Pulse-time Modulation (P.T.M.)

In P.T.M. the amplitude of each sample of the modulating wave is used to vary the time of occurrence of some feature of a pulse. For example, in one form of P.T.M. the time of occurrence of one edge of each pulse is varied; this results in assymmetric pulse-duration modulation (P.D.M.). In another form the modulating wave varies the times of occurrence of constant (usually short) duration pulses with respect to regularly recurring instants. If the modulating wave is used directly to vary the displacement of the pulse position, the system is termed pulse-position modulation (P.P.M.) but if the modulating signal is first passed through a circuit whose gain varies inversely with frequency the pulses are found to be modulated in frequency (P.F.M.).

The two forms of P.T.M. most applicable to combined vision-and-sound systems are P.P.M. and P.D.M. Fig. 4(*a*) shows part of a typical modulating wave. In P.P.M., illustrated in Fig. 4(*b*), the deviation in "position" (i.e. change of time of occurrence) of each constant-duration pulse, with reference to a regular recurring time datum, is proportional to the instantaneous a.f. signal amplitude; in the example shown, the pulse is delayed with reference to the datum when the instantaneous value of the a.f. signal is "positive" and is advanced for "negative" values. Fig. 4(*c*) illustrates

P.D.M., in which the duration of each pulse is determined by the instantaneous a.f. signal amplitude; a positive value of the a.f. signal lengthens the pulse and vice versa.

It will be appreciated that P.D.M. is closely related to P.P.M.; each pulse of a P.D.M. train has two edges either, or both,* of which may be modulated in position according to the instantaneous magnitude of the a.f. wave. P.P.M. may be converted to asymmetric P.D.M. (i.e. where only one edge of the pulse is modulated) by arranging that each position-modulated pulse initiates one edge (leading or trailing) of a further pulse whose other edge is initiated by a pulse that recurs regularly at the mean repetition frequency of the P.P.M. train and is timed so as to occur outside the interval (or "time slot") occupied by the position-modulated pulse. Similarly, P.D.M. may be converted to P.P.M. by deriving a train of constant-duration pulses each of which is coincident with the position-modulated edge of a duration-modulated pulse.

It is evident that both P.P.M. and P.D.M. involve a sampling process but, before considering the spectra corresponding to such forms of modulated-pulse train, it is important to know precisely the instants at which the a.f. modulating wave is sampled. If the a.f. wave is sampled at regularly recurring instants and the amplitude of each sample is then used to deviate the position of a pulse forming part of a P.P.M. train or used to control the timing of a pulse edge in a P.D.M. train, the system is then said to employ uniform (or periodic) sampling. However, if the effective instants of sampling coincide with the actual timings of the modulated pulses in P.P.M., or with the actual time-modulated edges of the modulated pulses in P.D.M., the system is then said to employ natural (or synchronous) sampling.

3.2.1. Pulse-position Modulation

It has been shown[6] that, with uniform sampling, the pulse-position modulation of extremely short pulses of unit area by a single tone results in a spectrum of the form:

$$F_u(t) = \frac{1}{T}\left[1 + 2\sum_{n=1}^{\infty} J_n\left\{n \cdot \frac{\omega_a}{\omega_r} \cdot \frac{2\pi d}{T}\right\} \cdot \cos n\omega_a t \,.\right]$$

$$+ \frac{2}{T} \cdot \sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty} J_n\left\{\left(m + n \cdot \frac{\omega_a}{\omega_r}\right) \cdot \frac{2\pi d}{T}\right\} \cdot \cos(m\omega_r + n\omega_a)\, t \,. \qquad (1)$$

where $\omega_a/2\pi$ is the frequency of the audio modulation (Hz)

$\omega_r/2\pi = 1/T$ is the mean pulse-recurrence frequency (Hz)

d is the peak pulse-position deviation (secs)

m and n are integer parameters defining the orders of the pulse-recurrence-frequency components and the audio-frequency components respectively.

With natural sampling:[6,7]

$$F_N(t) = \frac{1}{T}\left[1 + \frac{\omega_a}{\omega_r} \cdot \frac{2\pi d}{T} \cdot \cos \omega_a t\right]$$

$$+ \frac{2}{T}\sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty} \left(1 + \frac{n}{m} \cdot \frac{\omega_a}{\omega_r}\right) \cdot J_n\left(m \cdot \frac{2\pi d}{T}\right) \cdot \cos(m\omega_r + n\omega_a)t \qquad (2)$$

Figs. 5(*a*) and 5(*b*) illustrate the lower-frequency portions of typical spectra corresponding to single-tone, low-deviation P.P.M. with uniform and natural sampling respectively. It will be seen that both types of sampling are characterised by sets of sidebands, corresponding to the modulation frequency and its harmonics, associated with integer multiples of the pulse-recurrence frequency.

* In such a case the two edges are modulated in opposing senses; if they are equally modulated this results in symmetrical P.D.M.

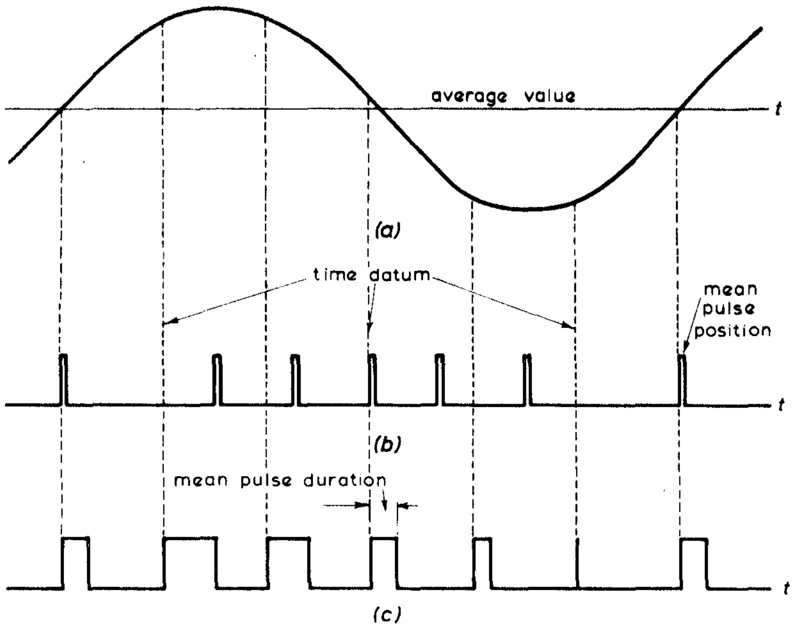

Fig. 4 - P.P.M. and P.D.M.

(*a*) Modulating wave
(*b*) Pulses position-modulated by (*a*)
(*c*) Pulses duration-modulated by (*a*)

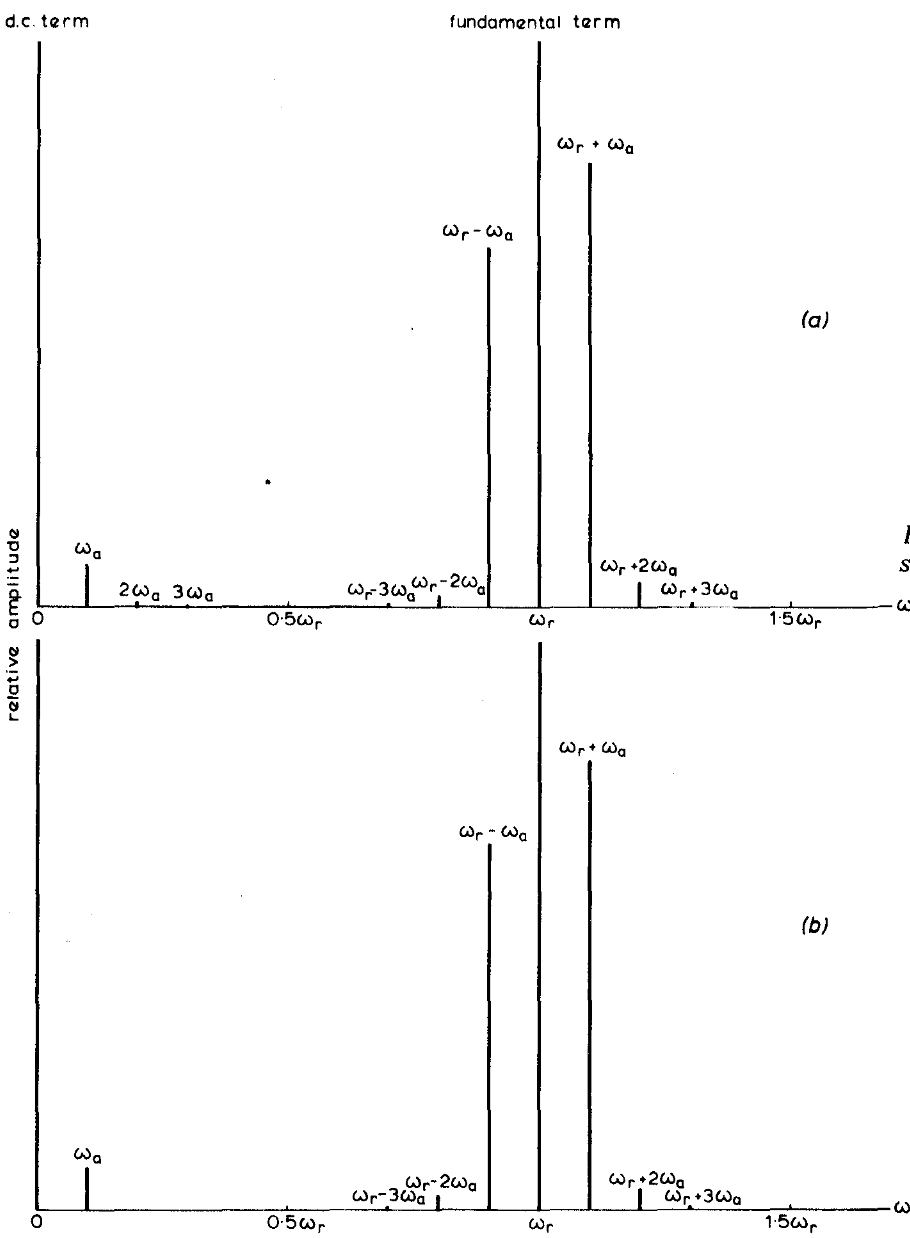

Fig. 5 - Lower frequency portions of spectra corresponding to P.P.M. with

(*a*) Uniform sampling
(*b*) Natural sampling

The only significant difference between the two spectra lies in the baseband components. With uniform sampling these consist of the modulation frequency and its harmonics; the amplitude of the baseband component representing the modulation is proportional to the ratio of modulating frequency to the pulse-recurrence frequency and the amplitudes of the baseband components representing the harmonics of the modulation frequency are approximately proportional to the square, cube, etc. of the above-mentioned ratio. The nature of the spectrum also indicates that, if more than one modulation component is present, beats are produced between the various modulating tones.

However, if the relative pulse-position deviation (d/T) is low (a necessary condition if the modulated pulses are to be included within a video waveform) the amplitudes of the unwanted baseband components, and the beats produced if several modulation components are present simultaneously, are small compared to the amplitude of the wanted sideband.

With natural sampling only one baseband component is produced and this occurs at the modulation frequency and its amplitude is proportional to the ratio of its frequency to the pulse-recurrence frequency. Two or more modulating signals result in an identical number of baseband components and beat-frequency signals are not produced.

The lower-sideband structures associated with the pulse-recurrence frequency are somewhat similar for both uniform and natural sampling. In both cases lower sidebands are generated which correspond to the modulating signal and its harmonics, and it is apparent that they can occur, for the higher modulation frequencies, within the band from zero to half the pulse-recurrence frequency. Further, simultaneous modulation by two or more tones results in lower sidebands of the carrier wave at pulse-recurrence frequency which correspond to beats between the modulation tones.

For a value of peak deviation typical of a P.P.M. signal suitable for combination with a 625-line video signal, the amplitude of the second-order lower sideband of the carrier wave at pulse-recurrence frequency (corresponding to the second harmonic of the modulation tone) is fairly small compared to that of the baseband component describing the modulation and is approximately proportional to the square of the deviation; the third and higher-order sidebands of the pulse carrier wave have very small amplitudes approximately proportional to the cube, fourth power etc., of the deviation, and the beats between modulation tones are negligible.

3.2.2. Pulse-duration Modulation

As in the case of pulse-position modulation a train of pulse-duration modulated pulses may be derived using either natural or uniform sampling. However, as indicated earlier, modulation may also be carried out either symmetrically or asymmetrically; in the former case the positions of both pulse edges are varied in opposite directions in accordance with the modulation and, in the latter case, the position of only one edge is varied.

With uniform sampling an asymmetrical P.D.M. pulse train may be expressed as:[6*]

$$F_u'(t) = \frac{2\delta}{T}\left[\sum_{n=1}^{\infty}\left\{\frac{J_n\left(n \cdot \frac{\omega_a}{\omega_r} \cdot \frac{2\pi\delta}{T}\right)}{n \cdot \frac{\omega_a}{\omega_r} \cdot \frac{2\pi\delta}{T}}\right\} . \sin n\omega_a t\right] + \frac{1}{\pi}\sum_{m=1}^{\infty}\left[\left\{\sum_{n=-\infty}^{\infty}\left(\frac{J_n\left(\overline{m + n \cdot \frac{\omega_a}{\omega_r}} \cdot \frac{2\pi\delta}{T}\right)}{\left(m + n \cdot \frac{\omega_a}{\omega_r}\right)}\right) . \sin(m\omega_r + n\omega_a)t\right\} - \frac{\sin m\omega_r t}{m}\right] \tag{3}$$

where δ is the peak positional deviation of the modulated pulse edge (secs). With natural sampling:[6,7*]

$$F_N'(t) = \frac{\delta}{T} . \sin \omega_a t + \frac{1}{\pi}\sum_{m=1}^{\infty} . \frac{1}{m}\left[\left\{\sum_{n=-\infty}^{\infty} J_n\left(m . \frac{2\pi\delta}{T}\right) . \sin(m\omega_r + n\omega_a)t\right\} - \sin m\omega_r t\right] \tag{4}$$

* A train of P.D.M. pulses may be regarded as the sum of a train of unmodulated pulses each having a duration greater than δ, together with a further train containing all information concerned with the modulation; the second pulse train alternates in polarity at the modulating frequency. In Equations (3) and (4) the terms describing the unmodulated pulses (i.e. a d.c. term and terms representing the pulse-recurrence frequency and its harmonics) have been discarded.

The corresponding expressions for symmetrical P.D.M. may be easily derived by regarding a train of pulses with both edges modulated as a combination of two asymmetrical P.D.M. pulse trains, the first having one edge modulated and the second consisting of a version of the first in which the time variable is reversed in sign.

As with P.P.M. the spectrum corresponding to uniform sampling and single-tone modulation includes baseband components at multiples of the modulating frequency. However, for low values of δ/T, the amplitude of the baseband component at modulation frequency is substantially independent of the modulation frequency and the amplitudes of those corresponding to harmonics of the modulating frequency are small; with several modulation frequencies present, low-amplitude beats can occur as in P.P.M. with uniform sampling. With natural sampling and single-tone modulation only one baseband component occurs at modulation frequency and its amplitude is independent of modulation frequency; two or more modulation tones produce corresponding baseband components but none corresponding to beats between the modulating tones are produced.

With low values of δ/T the lower sidebands of the pulse carrier are again fairly similar for both uniform and natural sampling; as in the case of P.P.M., these lower sidebands correspond to the modulating signal and its harmonics. However, unlike P.P.M. the amplitude of a sideband corresponding to the modulation or one of its harmonics is substantially independent of modulating frequency for both forms of sampling. Nevertheless, the amplitude of the second-order lower sideband of the pulse carrier is again fairly small compared with that of the baseband component describing the modulation tone and is approximately proportional to the square of δ/T; the third and higher-order lower sidebands have very small amplitudes approximately proportional to the cube, fourth-power, etc. of δ/T. Simultaneous modulation by two or more tones results in lower sidebands of the pulse carrier corresponding to beats between the modulating tones but, for low values of δ/T, their amplitudes are very small.

4. GENERATION AND DEMODULATION

4.1. P.A.M.

The generation of a P.A.M. pulse train and its demodulation have been outlined, in principle, in Section 3.1. The pulse train may be produced using a conventional form of sampling gate, driven by constant-amplitude pulses, to which the a.f. wave is applied. The demodulation of the P.A.M. signal is carried out, as already mentioned, by means of a suitable low-pass filter.

4.2. P.T.M.

It has already been pointed out that P.P.M. and P.D.M. are closely related and that one may be easily derived from the other. As a consequence it is necessary to discuss only the generation and demodulation of either P.P.M. or P.D.M. In the ensuing sections the generation and demodulation of P.P.M., with uniform and natural sampling respectively, are discussed.

4.2.1. P.P.M. With Uniform Sampling

The use of uniform sampling can provide a P.P.M. system free from unwanted distortion caused by intrusion into the baseband of sidebands of the pulse carrier corresponding to harmonics of the modulation. Techniques enabling such a result to be obtained involve the use of P.A.M. as an intermediate step in both the generation and demodulation of the signal. Fig. 6 illustrates the process of generation.

The a.f. wave (*a*) is first regularly sampled (at the pulse-recurrence frequency) at the instants t_1, t_2, t_3, etc. and the samples so obtained may then be applied to a "sample-and-hold" circuit in which each sample initiates a flat-topped pulse whose amplitude is equal to that of the sample and whose duration is equal to the interval between samples. The resulting waveform (*b*) is then compared, in instantaneous magnitude, with regularly recurring ramp waves, initiated at the time t_1, t_2, t_3, etc. and having durations equal to twice the peak deviation required in the output P.P.M. train, as shown in (*c*). At the instant at which the instantaneous magnitude of a ramp wave equals the magnitude of the waveform (*b*) one pulse of the P.P.M. train (*d*) is generated. It will be seen that the time-datum of each position-modulated pulse is delayed, with respect to the corresponding instant of sampling, by the peak deviation d of the P.P.M. train.

The process of demodulation is illustrated in Fig. 7; for convenience the input (*a*) consists of the P.P.M. train of Fig. 6(*d*). Regularly recurring ramp waves (*b*), each having a duration not less than twice the peak positional deviation of the P.P.M. pulses (i.e. $2d$), are sampled by the input pulses. As a result each input pulse produces a sample whose amplitude is linearly related to its deviation; this results in a further train of pulses modulated in both position and amplitude. The position modulation may now be removed by applying the pulses obtained by sampling the ramp waves to a "hold" circuit which, as in the case of Fig. 6(*b*), results in a series of flat-topped pulses (*c*) initiated by the samples and having amplitudes directly related to the amplitudes of the samples. As before the duration of each flat-topped pulse is equal to the interval between the initiating sample

and its successor; however, in this case the samples are not spaced regularly. The waveform (*c*) may now be sampled by regularly recurrent pulses having the same recurrence frequency as the P.P.M. train (i.e. a period equal to $t_2 - t_1$) but so phased that sampling cannot occur within the intervals corresponding to the deviation range of the P.P.M. pulses; this process is shown in (*c*) where the sampling occurs at the instants $t_1 + \theta$, $t_2 + \theta$, etc. The result is to form a train of regularly recurring pulses which are amplitude modulated by the a.f. wave (i.e. P.A.M. train). The a.f. wave may then be recovered, free from distortion, by applying the P.A.M. pulses to a suitable low-pass filter.

From Figs. 6 and 7 it will be seen that the a.f. wave has suffered a delay equal to or greater than twice the peak deviation of the P.P.M. pulses.

4.2.2. P.P.M. With Natural Sampling

The generation of P.P.M. with natural sampling is illustrated in Fig. 8; in general, it is similar to the equivalent process used for uniform sampling. However, the a.f. wave (*a*) is now compared directly with regularly recurring ramp waves; no intermediate sampling process is used. As shown in (*b*), the ramp waves, whose mid-points occur at T_1, T_2, T_3, etc., recur at a frequency equal to the mean recurrence frequency required for the output P.P.M. train. The instantaneous magnitude of each ramp wave is compared with that of the a.f. wave and one pulse of the output P.P.M. train (*c*) is generated when they are equal. It will be seen that the time-datum of output pulses occurs at T_1, T_2, T_3, etc., and that the deviation of each pulse describes the instantaneous magnitude of the a.f. wave at the instant at which the pulse is produced.

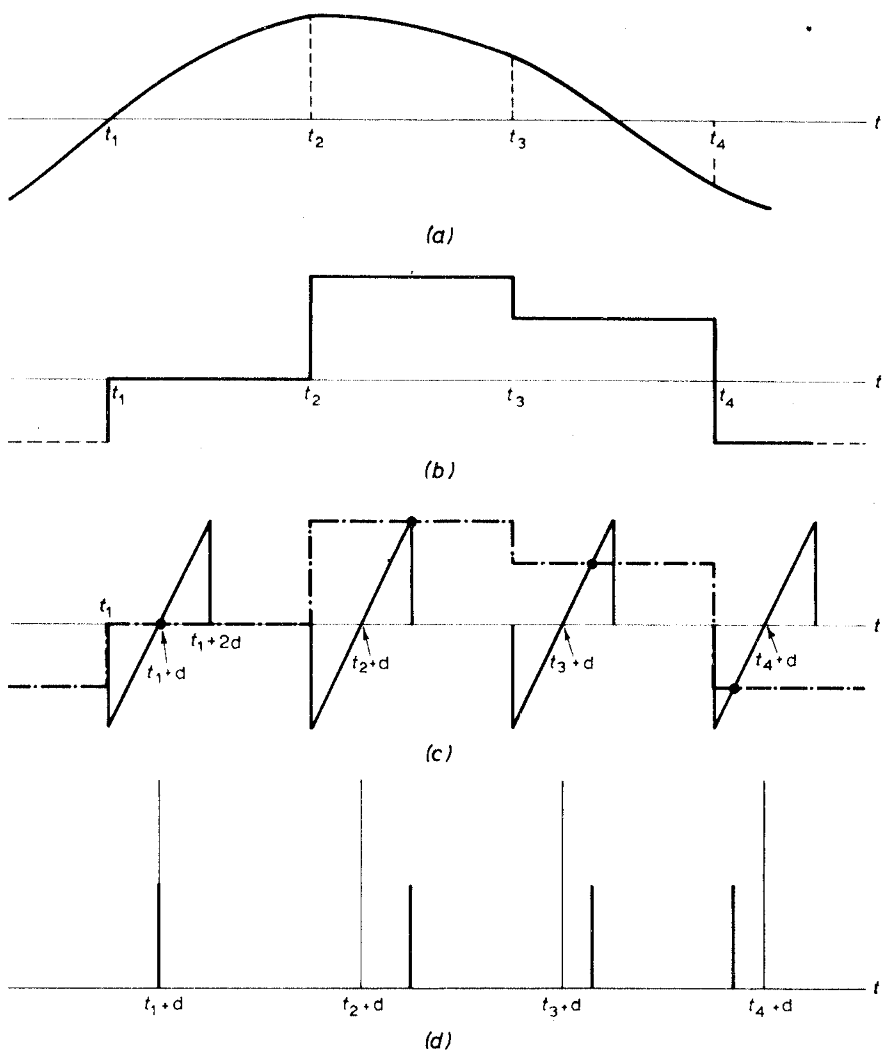

Fig. 6 - Generation of P.P.M. with uniform sampling
(*a*) Modulating a.f. wave (*b*) Result of "sampling-and-holding" (*a*)
(*c*) Comparison of (*b*) with regular ramp waves (*d*) Output pulse train

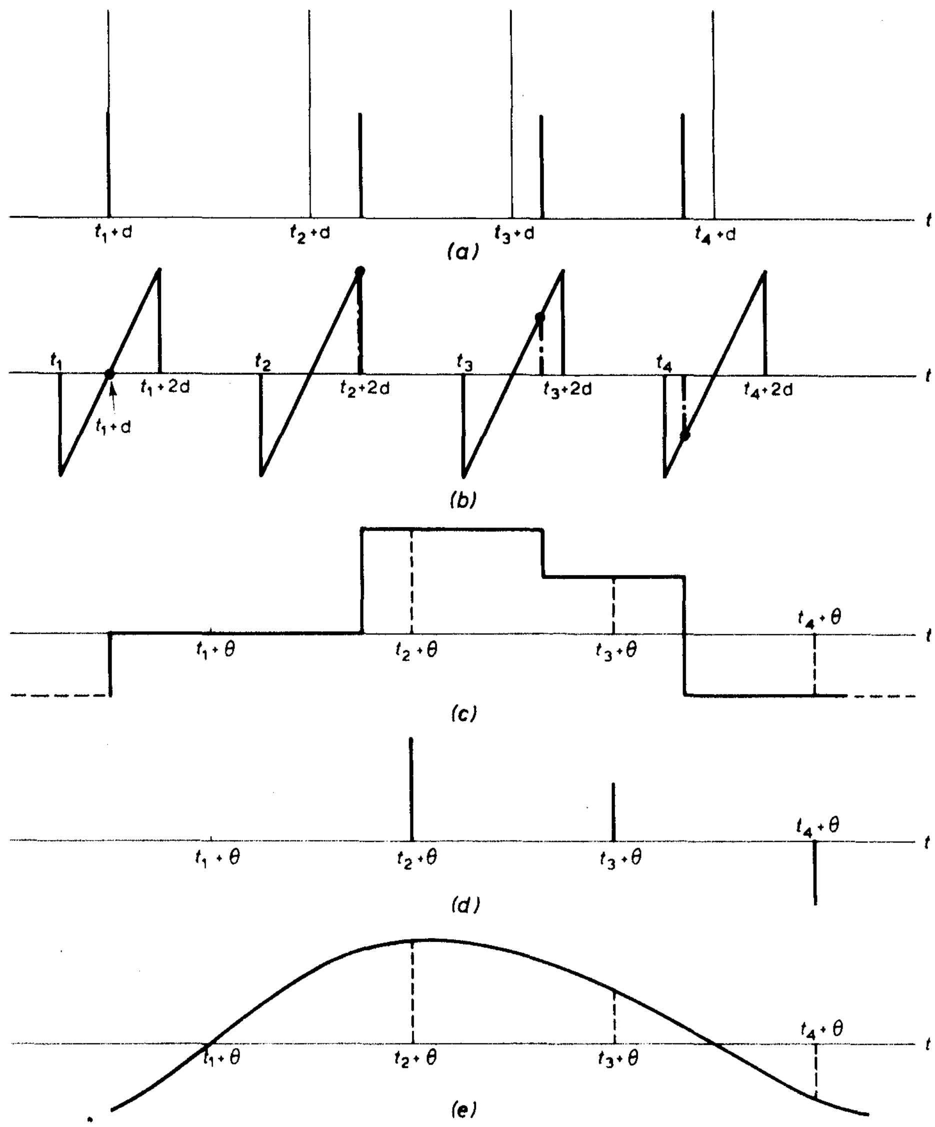

Fig. 7 - Demodulation of P.P.M. with uniform sampling
(*a*) P.P.M. pulse train with uniform sampling (*b*) Pulses of (*a*) sampling regular ramp waves
(*c*) Samples of (*b*) after "holding" (*d*) Samples produced by regularly sampling (*c*) (*e*) Output a.f. wave

Demodulation of P.P.M. with natural sampling may be performed in several ways. One method uses P.D.M. as an intermediate step. As mentioned in Section 3.2, a P.P.M. pulse train may be used to form asymmetric P.D.M. by causing each input pulse to initiate one edge (leading or trailing) of a further pulse whose other edge is initiated by one pulse of a regular train repeating at the mean pulse-recurrence frequency of the input P.P.M. train. The P.D.M. pulse train thus formed is characterised by natural sampling and a study of Equation 4 reveals that the corresponding spectrum contains only baseband components describing the modulating wave. Thus, the wanted modulation may be derived by passing the P.D.M. signal through a suitable low-pass filter. However, as mentioned earlier, unwanted "inversions" of modulation frequency signals, due to lower sidebands of the carrier wave at pulse-recurrence frequency, may intrude into the passband of the filter, causing distortion.

A train of P.P.M. pulses with natural sampling and parameters typical of a signal suitable for combination with a video signal (d = 1·5 μs, T = 64 μs and f_r = 15,625 Hz) may be converted into a corresponding P.D.M. train (with δ = 1·5 μs) whose spectrum contains a lower-sideband of the carrier at pulse-recurrence frequency having an amplitude approximately -29 dB with respect to that of the baseband component representing the modulation tone and independent of modulation frequency.

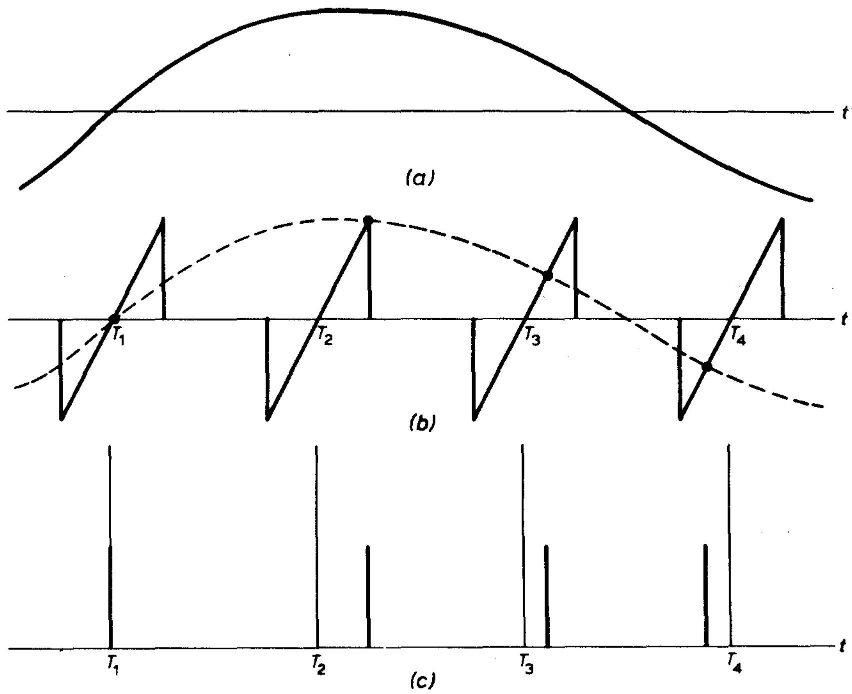

Fig. 8 - Generation of P.P.M. with natural sampling
(*a*) Modulating a.f. wave (*b*) Comparison of (*a*) with regular ramp waves (*c*) Output pulse train

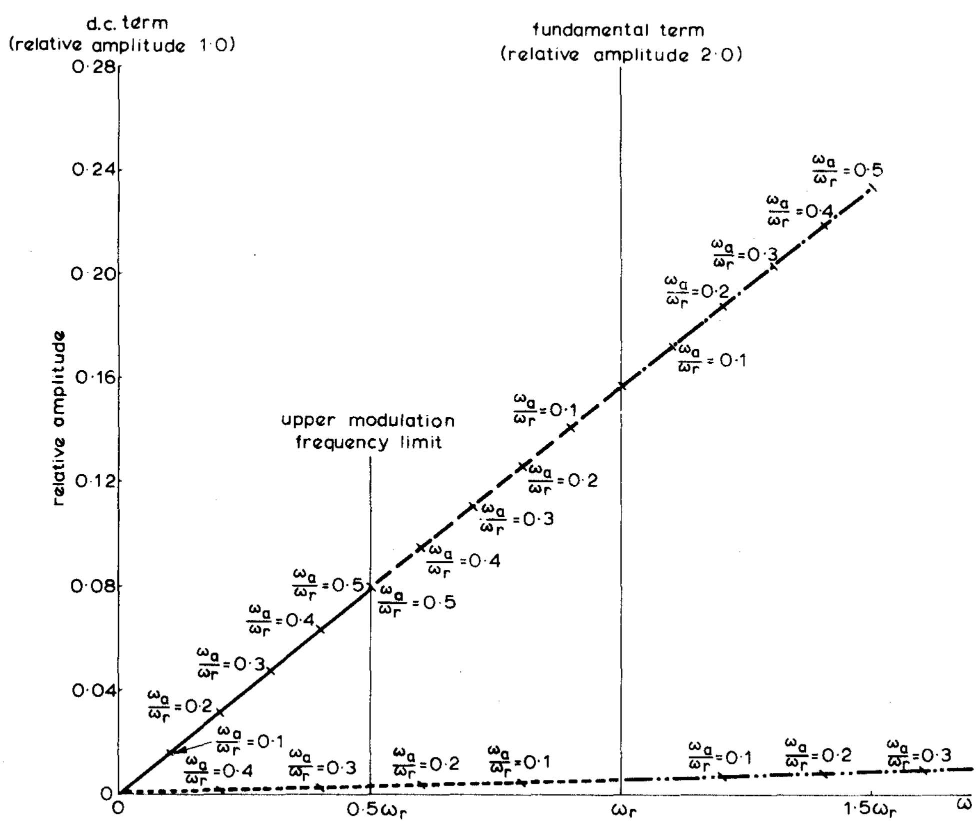

Fig. 9 - Amplitudes, as functions of modulating frequency, of lower frequency spectral components of typical P.P.M. train with natural sampling

——— baseband component ——— first-order lower sideband of p.r.f.
—·—·— first-order upper sideband of p.r.f. ------ second-order lower sideband of p.r.f.
—··—··— second-order upper sideband of p.r.f.

As the modulation frequency is raised from $f_r/4$* to $f_r/2$, whilst the peak deviation δ remains constant, the frequency $(\omega_r - 2\omega_a)/2\pi$, see Fig. 5, of the lower sideband of the carrier corresponding to the second harmonic of the modulation falls from $f_r/2$ to zero, the amplitude remaining constant.

A second method of demodulating P.P.M. with natural sampling is apparent from consideration of the signal spectrum discussed in Section 3.2.1. Fig. 9 illustrates certain properties of a typical spectrum and shows how the amplitudes of the baseband components and the first and second order sidebands of the p.r.f. vary as a function of the ratio between the modulation frequency and the pulse-recurrence frequency. Fig. 9 also shows that the amplitude of the second-order lower sideband of the signal at pulse-recurrence frequency falls, substantially linearly, as the modulation frequency increases; this unwanted component does not intrude into the wanted band for modulation frequencies less than half the upper limit, $f_r/4$, and the ratio of the amplitude of the unwanted inversion to that of the wanted modulation is a maximum at this particular frequency. It is apparent that the modulation, described by the baseband component may be derived by applying the modulated pulse train to a corrected network whose response, as a function of frequency, falls at 6 dB per octave from the lowest modulation frequency to half the pulse-recurrence frequency and is zero for higher frequencies. However, the effect of the correcting network is progressively to reduce the amplitude of the wanted component as its frequency is increased whilst progressively increasing the amplitude of the inversion (whose frequency falls as the modulation frequency rises from a quarter to half the pulse-repetion frequency).

Fig. 10 illustrates this effect of the correcting network. At the input to the network the wanted modulation has a frequency characteristic rising at 6 dB per octave. Modulation-frequency signals in the frequency range between $f_r/4$ and $f_r/2$ (Fig. 10 (*a*)) give rise to inversions (Fig. 10(*b*)) between $f_r/2$ and zero; the frequency characteristic of the inversion also rises at 6 dB per octave. The characteristics of the modulation and its inversion are shown as solid lines in the figure. It has been assumed that the correcting network has unity gain at $f_r/2$ and its effects upon the modulation and the inversion are shown as dotted lines; their amplitudes are now independent of modulation frequency. However, modulation at a frequency $f_r/4$ is doubled in amplitude while the amplitude of its inversion, at $f_r/2$ is unchanged. Thus, at the output of the correcting network, the ratio of the amplitude of the wanted modulation to that of its inversion is independent of modulation frequency.

* $f_r = \omega_r/2\pi$, the pulse-recurrence frequency.

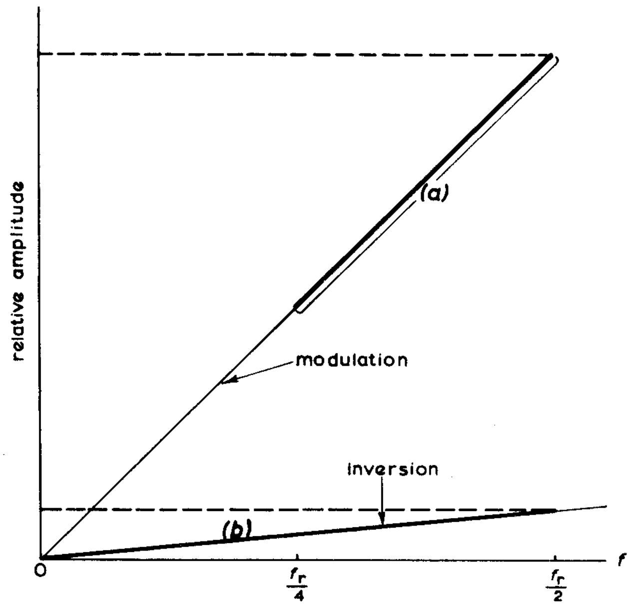

Fig. 10 - Effect of correcting network on P.P.M. with natural sampling

(*a*) Range of modulating frequency giving rise to inversion due to:
(*b*) Second order lower sideband of p.r.f.
——— Input to correcting network
– – – – – – Output from correcting network

For the P.P.M. system-parameters described previously (i.e. d = 1·5 μs, T = 64 μs and f_r = 15,625 Hz) the ratio of the inversion amplitude to that of the wanted modulation is approximately -29 dB at the output of the correcting network and independent of modulation frequency; this is a result in agreement with that obtained using P.D.M. as an intermediate step.

A third method of demodulating P.P.M. with natural sampling is illustrated in Fig. 11. As in the case of P.P.M. with uniform sampling, Fig. 7, the input train (*a*) samples regularly recurring ramp waves (*b*) so phased that each ramp embraces the deviation range of the input pulses. However, in this case, the resulting train of samples, modulated in both position and amplitude, is applied directly to a suitable low-pass filter.

The operation of such a demodulator may be understood by considering the spectrum of the P.P.M. pulse train and that of the ramp wave; for the sake of simplicity the ramp wave may be considered as part of a sine wave at pulse recurrence frequency. Fig. 12(*a*) shows qualitatively the lower-frequency components* of the spectrum of P.P.M. with natural sampling together with a spectral line at the pulse-repetition frequency f_r representing the ramp wave; in Fig. 12(*a*) it has been assumed that the modulation frequency f_a lies between $f_r/4$ and $f_r/2$.

* Sidebands of the carrier wave at pulse-recurrence frequency of higher order than 2 have been neglected.

Fig. 12(b) illustrates qualitatively the output of the ramp demodulator obtained by multiplying the sine wave by the P.P.M. pulses; only components lower in frequency than f_r are shown. It will be seen that several components of the P.P.M. spectrum, after beating with the demodulating sine wave, cause output components to appear. The principle output component occurs, as desired, at f_a but unwanted components also appear at $(f_r - 2f_a)$, $(f_r - f_a)$ and $2f_a$. Applying the output of the ramp demodulator to a low-pass filter removes $(f_r - f_a)$ and either $(f_r - 2f_a)$ or $2f_a$. If the modulation frequency f_a were lower than $f_r/4$ the $(f_r - 2f_a)$ inversion component would be rejected but the second-harmonic distortion component $2f_a$ would be passed to the output of the filter. As f_a lies between $f_r/4$ and $f_r/2$ the second-harmonic component is rejected but, in this case, the inversion component appears at the filter output.

Analysis based upon the use of a sine wave representing the ramp wave shows that, for the P.P.M. system parameters described previously, the amplitudes of the wanted modulation, the second harmonic and the inversion components behave, as functions of frequency, in the manner shown in Fig. 13.

It will be seen that, for the three methods of demodulation for P.P.M. with natural sampling, distortion terms are produced. In all cases the level of the significant distortion is approximately proportional to the square of the pulse deviation, which itself directly determines the output modulation-signal amplitude; for a modulation signal causing half the peak pulse deviation, the relative level of each of the two distortion contributions is halved. The value of d (1·5 μs) assumed in the example quoted in this section corresponds to the peak value of the deviation. As the average television-sound signal level is appreciably lower than the peak, the average level of distortion would be significantly less than that indicated.

Further, as the distortion introduced by each form of demodulator may be determined accurately, it is possible* to devise correction, at the generator of the modulated-pulse train, and thus to eliminate distortion from the demodulator output. Where all forms of demodulator are used with the same pulse train, a compromise correction may be devised which significantly reduces output distortion from all demodulators.

* As pointed out by G.D. Monteath.

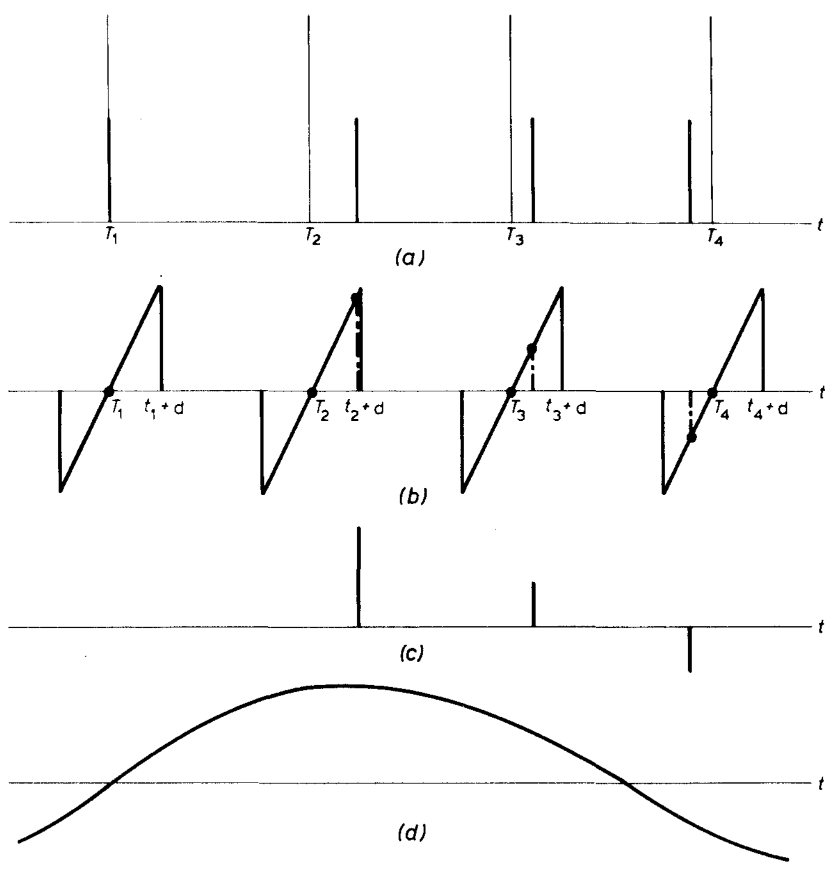

Fig. 11 - Demodulation of P.P.M. with natural sampling, using ramp waves
(a) P.P.M. pulse train with natural sampling (b) Regular ramp waves sampled by (a)
(c) Samples produced as a result of (b) (d) Output a.f. wave

5. SIGNAL-TO-NOISE RATIO

In a modulated pulse train, information describing the modulation is transmitted during only a small interval during each pulse-recurrence period. Hence, in order to obtain maximum signal-to-noise ratio from such a system it is necessary to prevent noise reaching the demodulator except during the above-mentioned small intervals. The following sub-sections, describing the signal-to-noise characteristics of P.A.M. and P.T.M., include outlines of the means used to achieve this.

The form of noise considered consists of fluctuation noise, as produced by thermal agitation in the input circuits of a receiver. No attempt is made to assess the effects of impulsive noise.

5.1. P.A.M.

In a P.A.M. system the maximum signal-to-noise ratio is obtained by gating the signal so as to remove all noise input to the demodulator during the constant intervals between pulses; such gating also removes any unwanted signals (e.g. video signals) occurring during these intervals.

If V_p is the peak voltage excursion of the pulse amplitude at the input to the low-pass filter used for demodulation,*

τ is the pulse duration (secs),

T is the pulse-recurrence period (secs),

* For 100% modulation, V_p is equal to the unmodulated-pulse amplitude.

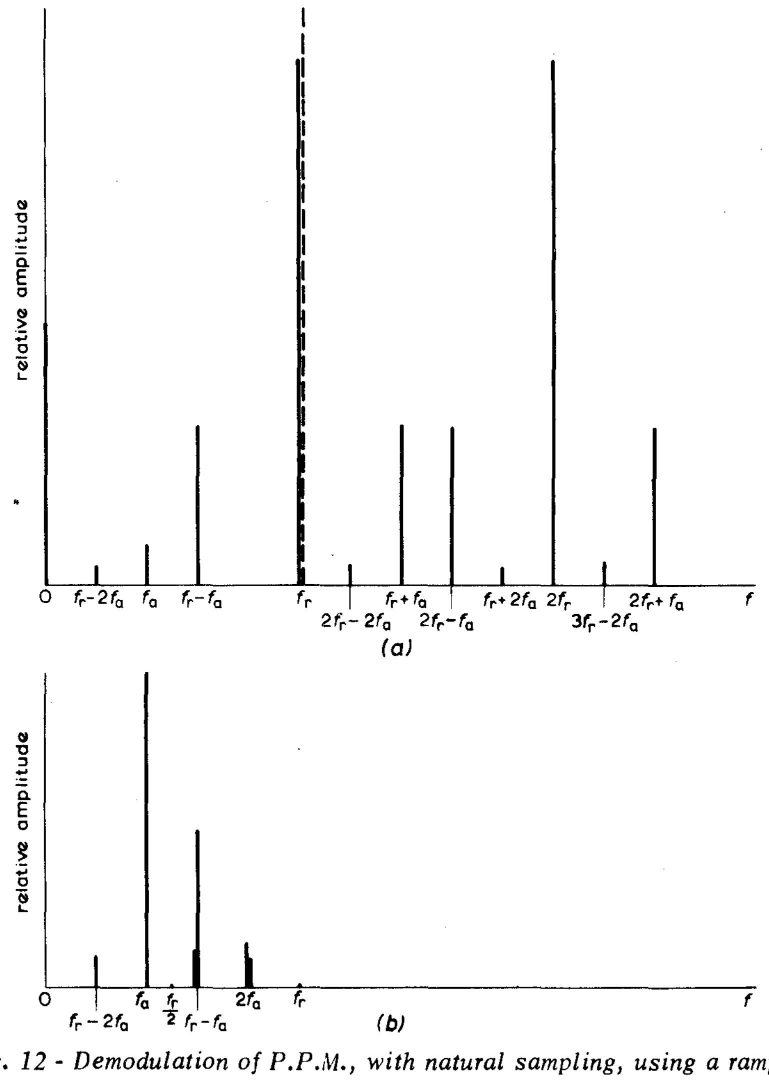

Fig. 12 - Demodulation of P.P.M., with natural sampling, using a ramp wave

(a) Spectral components of P.P.M. pulses with natural sampling, together with component representing ramp wave ($f_r/4 < f_a < f_r/2$)

(b) Components of ramp-demodulator output having frequencies less than f_r ($f_r/4 < f_a < f_r/2$)

-------- component representing ramp wave ———— components of P.P.M. pulses

V_N is the r.m.s. noise voltage at the filter input during the pulse interval,

f_o is the effective video bandwidth of the signal and noise applied to the filter (Hz)

and $f_a = 1/2T$ is the a.f. bandwidth (Hz),

the peak signal, V_s, obtained by filtering the pulses is

$$V_s = V_P \cdot \frac{\tau}{T}$$

The average noise power, $\overline{v_N^2}$, fed to the filter is:

$$\overline{v_N^2} = V_N^2 \cdot \frac{\tau}{T}$$

thus the average noise-power output, V^2, from the filter is:

$$V^2 = V_N^2 \cdot \frac{\tau}{T} \cdot \frac{f_a}{f_o}$$

$$= V_N^2 \cdot \frac{\tau}{T} \cdot \frac{1}{2Tf_o}$$

whence the r.m.s. noise-output voltage, V, is:

$$V = \frac{V_N}{T} \cdot \sqrt{\frac{\tau}{2f_o}}$$

thus:

$$\frac{V_s}{V} = V_P \cdot \frac{\tau}{T} \cdot \frac{T}{V_N} \cdot \sqrt{\frac{2f_o}{\tau}}$$

and the ratio, $R_{P.A.M.}$, of the r.m.s. output-signal voltage to the r.m.s. output noise voltage:

$$R_{P.A.M.} = \frac{V_P}{V_N} \cdot \sqrt{\tau f_o} \; \ldots\ldots \quad (5)$$

Certain deductions may be drawn from Equation (5). First, provided that the video bandwidth, f_o, is sufficient to prevent significant distortion of the modulated-pulse shape, the value of f_o does not affect the output signal-to-noise ratio; an increase of f_o is compensated by a corresponding increase in V_N (which is proportional to $\sqrt{f_o}$). If, however, f_o were reduced to a value such as to affect the pulse amplitude and duration, then:

$$V_p \propto f_o, \; \tau \propto \frac{1}{f_o}$$

and:

$$V_N \propto \sqrt{f_o}$$

whence, from Equation 5, $R_{P.A.M.} \propto \sqrt{f_o}$.

Thus, it is desirable to ensure that the form of pulse used in P.A.M. has a spectrum confined well within the video bandwidth available.

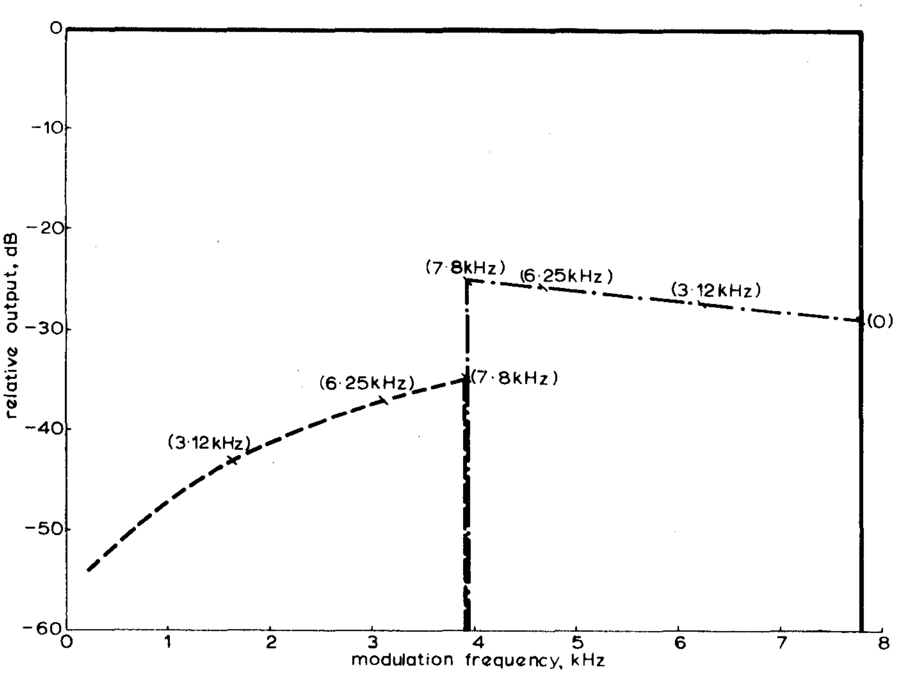

Fig. 13 - Output components of ramp demodulator of typical P.P.M., with natural sampling, as functions of modulating frequency

———— modulation component; – – – – – unwanted second harmonic component; —·—·— unwanted inversion component; (   ) frequency of unwanted component

5.2. P.T.M.

In a P.T.M. system gating arrangements may be used, as in P.A.M., in order to remove unwanted signals and noise occurring during the intervals between pulses; in this case, however, the gate must pass the required signal, together with noise for all positions of the modulated pulse edges. In P.P.M. and asymmetrical P.D.M. systems all essential information concerning the modulation is indicated by the position of one edge of each pulse;* further, pulse amplitude is not a parameter describing the modulation. Thus, for the above-mentioned P.T.M. systems, the circuit arrangements prior to the demodulator should incorporate amplitude limiting (or slicing) together with gating so as to restrict the noise appearing at the demodulator input to that occurring during only a small portion of the rise (or fall) time of the pulse edge carrying modulation.

* Only in the case of symmetrical P.D.M. is it necessary to pass both edges to the demodulator.

Fig. 14 illustrates this procedure. Fig. 14(*a*) shows a time-modulated pulse accompanied by random fluctuation noise. Noise accompanying all low-slope regions of the pulse (e.g. the baseline and the peak) is removed, by limiting, to form the pulse shown in Fig. 14(*b*).

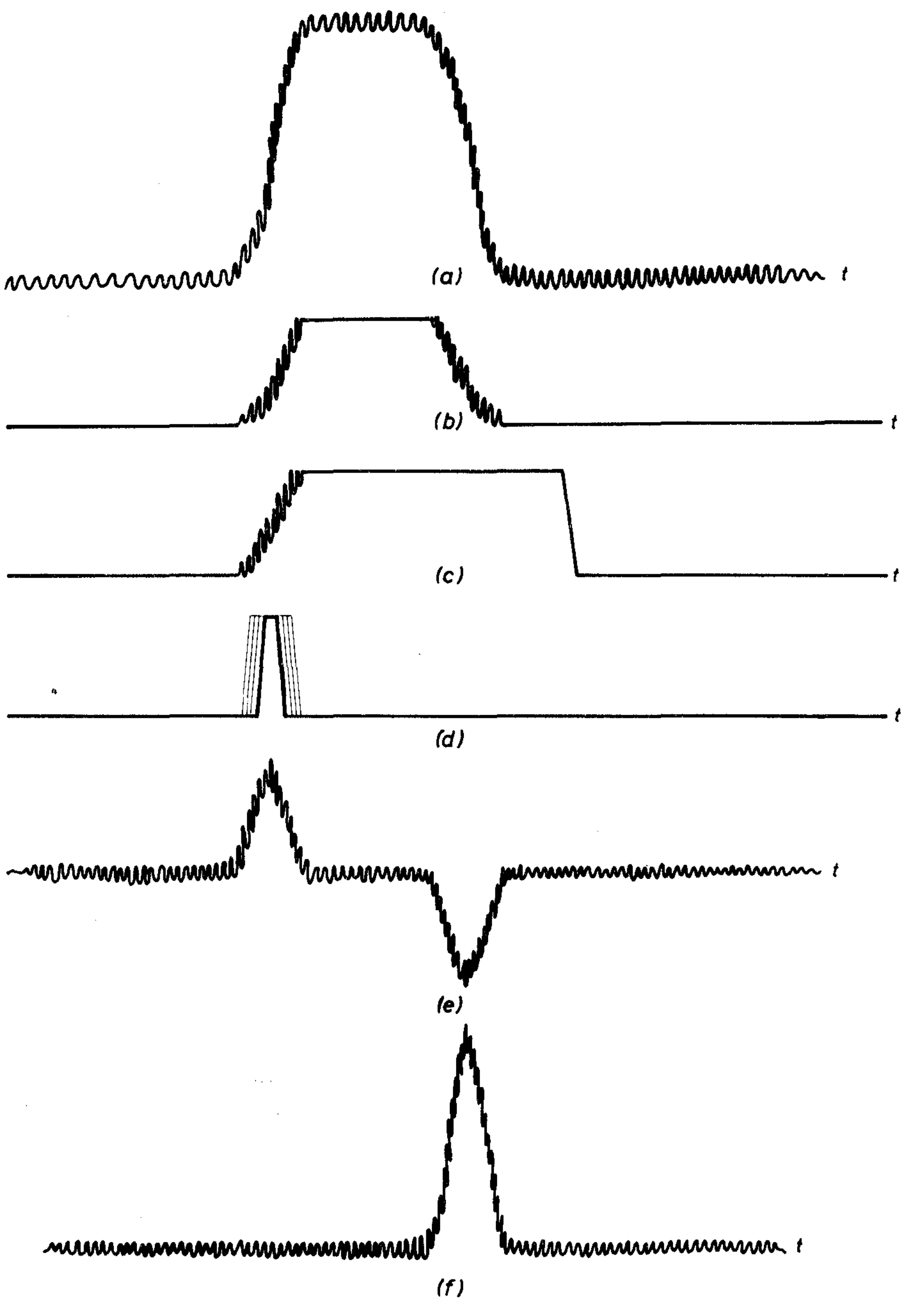

Fig. 14 - Minimising the effects of noise prior to demodulation of P.T.M.

(*a*) P.T.M. pulse with accompanying noise (*b*) Pulse produced by suitably limiting (*a*)
(*c*) Asymmetric P.D.M. pulse with noise eliminated from unmodulated edge (*d*) Short, constant duration, P.P.M. pulse
(*e*) Narrow pulses derived from edges of input pulse (*f*) Pulse produced by combining pulses of (*e*)

In the case of asymmetric P.D.M., a further modulated pulse may be derived, one edge being formed from the modulated edge of the pulse shown in Fig. 14(*b*) and the other, unmodulated edge, being formed using a regularly recurring pulse derived from an oscillator that is locked, by flywheel means, to the unmodulated edge of the pulse shown in Fig. 14(*a*). The derived pulse, illustrated in Fig. 14(*c*), is then passed to the demodulator.

For P.P.M. the corresponding process consists of using the pulse obtained after limiting to form a short modulated pulse, shown in Fig. 14(*d*), having a constant duration and a position that is perturbed by the noise accompanying only one (say the leading) edge of the pulse shown in Fig. 14(*b*);* this pulse is then utilised by the demodulator.

However, if the input P.P.M. pulses have durations such that the noise accompanying each leading edge is substantially uncorrelated with that accompanying each trailing edge, a more efficient process may be used. In such a case two narrow pulses, as shown in Fig. 14(*e*), may be formed (one from the leading edge of each input pulse and one from the trailing edge) by, for example, differentiation. The earlier of the two pulses may then be delayed so as to coincide with the later pulse and combined with it to form a pulse of double amplitude which is illustrated in Fig. 14(*f*); the associated increase in noise magnitude will, however, be only 3 dB. The pulse of Fig. 14(*f*) may now be limited and used to generate a pulse of the form shown in Fig. 14(*d*) which is utilised, as before, by the demodulator.

The signal-to-noise ratio obtained at the output of the demodulator for either asymmetric P.D.M. or for P.P.M. (assuming that only one edge of each pulse is used) may be derived quite simply.**

If V_N is the r.m.s. noise accompanying input pulses (volts),

and S is the slope, prior to limiting, of the pulse edge at the slicing level (V/s),

then the r.m.s. edge-timing error ϵ, due to noise, is[3,8]

$$\epsilon = V_N/S$$

If the peak signal output Vs from the demodulator is defined as:

$$Vs = \lambda \,.\, d \text{ (for P.P.M.)}$$

$$\text{or} = \lambda \,.\, \delta \text{ (for P.D.M.)}$$

* This may be carried out using a relaxation oscillator triggered by the pulse of Fig. 14(*b*).

** A more general analysis, based upon that given in Reference 3, is given in Appendix I.

where λ is a constant,

then the r.m.s. noise output V_N' resulting from an edge-timing error ϵ is:

$$V_N' = \lambda \,.\, V_N/S$$

and, for P.P.M., the r.m.s. signal to r.m.s. noise ratio, $R_{P.P.M.}$, at the demodulator output is:

$$R_{P.P.M.} = \frac{S \,.\, d}{V_N \,.\, \sqrt{2}} \quad \ldots\ldots \quad (6)$$

Correspondingly:

$$R_{P.D.M.} = \frac{S \,.\, \delta}{V_N \,.\, \sqrt{2}} \quad \ldots\ldots \quad (6a)$$

If, however, the noise accompanying the two edges of each P.P.M. pulse is uncorrelated, the process outlined in Figs. 14(*d*), (*e*) and (*f*) results in an improvement of 3 dB whence:

$$R'_{P.P.M.} = \frac{Sd}{V_N} \quad \ldots\ldots \quad (7)$$

Equations (6), (6*a*) and (7) show that, in each case, the signal-to-noise ratio is directly proportional to S. If the pulse shape, at the input to the limiter, is determined by the bandwidth of the circuit preceding the limiter, S is linearly related to the circuit bandwidth. In such circumstances, increasing this bandwidth raises V_N according to a square root law; as a result the signal-to-noise ratio is proportional to the square root of the bandwidth prior to limiting.

It will be appreciated that if the signal-to-noise ratio prior to the limiter is below a certain value, it may not be possible to prevent, by amplitude limiting, noise which occurs during low-slope regions of the pulse waveform from reaching the demodulator. As the input signal-to-noise ratio is reduced and the critical or "threshold" value reached, the signal-to-noise ratio at the demodulator output falls catastrophically.

5.3. Comparisons With A.M. and F.M.

The signal-to-noise performances of pulse-modulation systems may readily be compared with those of conventional A.M. and F.M. systems. In the ensuing discussion certain assumptions have been made with regard to the parameters which could be considered to be representative of typical television-sound systems; for example, it has been assumed that, in all cases, the a.f. bandwidth is 7·5 kHz. Further, it has been assumed that each of the sound systems is associated with the same television system whose vision-signal parameters

conform in all respects except one, with Standard I; the exception is the value of video bandwidth and it has been assumed that the effective bandwidth at the output of the vision detector of the receiver is 5 MHz.

5.3.1. A.M.

The signal-to-noise ratio of an A.M. television-sound system is related to that of the television system by several factors. These are: the ratio of the video bandwidth to the a.f. bandwidth, the forms of transmission used (i.e. whether double or asymmetric sideband), the ratio of the normalised peak-to-peak amplitudes of the modulating signals and the ratio of the peak carrier powers. Table 1 shows how these factors affect the ratio of $R_{A.M.}$ to R_V where $R_{A.M.}$ is the ratio of r.m.s. sound signal at 100% modulation to r.m.s. sound noise and R_V is the ratio of picture signal at white to r.m.s. video noise.

TABLE 1

Factor	Effect on $R_{A.M.}/R_V$ dB
Bandwidth ratio (video: 5 MHz a.f.: 7·5 kHz)	+28
Form of transmission (vision asymmetric, sound double side-band)	+ 3
Ratio of peak-to-peak amplitudes of modulating signals	+ 5
Peak carrier power ratio*	0
Correction of a.f. signal level to give r.m.s. value	- 9
TOTAL	+27

From Table 1 it will be seen that the ratio of the two signal-to-noise ratios, $R_{A.M.}/R_V$, for the parameters used has a value of +27 dB.

5.3.2. F.M.

The signal-to-noise performance of an F.M. television-sound system is very dependent upon the type of reception used. If it is assumed that the sound channel of the receiver utilises only the sound carrier (i.e. a "separate" sound channel) then the performance, $R_{F.M.}$, may be derived from that of the corresponding A.M. system having the same unmodulated carrier power by the well-known relation**

$$\frac{R_{F.M.}}{R_{A.M.}} = \sqrt{3 \cdot \frac{f_d}{f_a}} \qquad (8)$$

where f_d is the frequency deviation corresponding to 100% modulation.

If f_d has the value 50 kHz and the ratio of the unmodulated carrier power to the peak vision power is -7 dB (as in Standard I) then the ratio $R_{F.M.}/R_{A.M.}$ has the value +20 dB.

If, however, the receiver sound channel employs the intercarrier method, in which the F.M. sound carrier is heterodyned with the vision carrier so as to produce a further F.M. carrier whose frequency is independent of receiver tuning, the signal-to-noise performance of the sound system now depends upon the signal-to-noise ratios, at the receiver input, of both the sound and vision carriers.[9] The vision carrier signal-to-noise ratio depends upon the vision modulation characteristics of the television system and upon the actual picture-signal level; for Standard I, the vision-carrier signal-to-noise ratio, at the receiver input, for black level is some 12 dB greater than for white level.

It can be shown*** that the signal-to-noise ratio, $R'_{F.M.}$, obtained from a sound channel employing the intercarrier principle is related to that given by an ideal sound receiver, $R_{F.M.}$, by:

$$\frac{R'_{F.M.}}{R_{F.M.}} = \sqrt{1 + K^2} \qquad (9)$$

where K is the r.m.s. value of the ratio of the sound-carrier voltage (assumed constant) to the vision-carrier voltage at the receiver input.

For the television standards assumed, and a vision signal corresponding to white, K has a value of 2·05. Thus, in this case:

$$\frac{R'_{F.M.}}{R_{F.M.}} = -7{\cdot}5 \text{ dB}$$

whence $\dfrac{R'_{F.M.}}{R_{A.M.}} = +12{\cdot}5$ dB approximately.

5.3.3. P.A.M.

Equation 5 of Section 5.1 permits the signal-to-noise performance of a P.A.M. system to be compared with that of A.M. In the following comparison it has been assured that the P.A.M. pulse train consists of constant-duration pulses located in line-synchronizing intervals, in the manner shown in Fig. 2, and has the following parameters:

V_p equal to half the excursion from sync level to white level (i.e. 100% P.A.M. causes the

* Assuming the ratio of unmodulated sound-carrier power to peak vision-carrier power is -6 dB (as in Standard A).

** This neglects pre-emphasis and de-emphasis.

*** See Appendix II.

peak pulse amplitude to equal white level), and τ equal to 3 μs (a suitable value for Standard I).

Thus:

$$\frac{V_p}{V_N} = R_V \cdot \tfrac{1}{2} \cdot \frac{E_{sw}}{E_{bw}}$$

where E_{sw} is the signal excursion from sync level to white level and E_{bw} is the signal excursion from black level to white level.

Whence, from Equation (5):

$$R_{P.A.M.} = R_V \cdot \tfrac{1}{2} \cdot \frac{E_{sw}}{E_{bw}} \cdot \sqrt{\tau f_0}$$

therefore:

$$\frac{R_{P.A.M.}}{R_V} = \tfrac{1}{2} \cdot \frac{E_{sw}}{E_{bw}} \cdot \sqrt{\tau f_0} = +9 \text{ dB}$$

Thus, using the 27 dB figure from Table 1

$$\frac{R_{P.A.M.}}{R_{A.M.}} = -18 \text{ dB}$$

5.3.4. P.T.M.

The signal-to-noise performances, above threshold, of P.P.M. and P.D.M. for pulse parameters suitable for use with the television system assumed may be readily calculated using Equations (6), (6*a*) and (7) of Section 5.2.

For a P.P.M. (or P.D.M.) pulse train in which each pulse is located within a line-synchronizing interval, as illustrated in Fig. 2, the following parameters are convenient:

V_p equal to the excursion from sync level to white level

d (or δ) equal to 1·5 μs

S equal to the maximum slope of the edge of a sine-squared pulse whose peak magnitude is V_p and whose half-magnitude duration, x, is 0·2 μs.

Thus:

$$S = \frac{\pi}{2x} \cdot V_p$$

In this case:

$$\frac{V_p}{V_N} = R_V \cdot \frac{E_{sw}}{E_{bw}}$$

Assuming "single-edge" demodulation of P.P.M. (P.D.M. gives the same result) we have, from Equation 6:

$$R_{P.P.M.} = R_V \cdot \frac{E_{sw}}{E_{bw}} \cdot \frac{\pi \cdot d}{2x\sqrt{2}}$$

$$\frac{R_{P.P.M.}}{R_V} = +22 \text{ dB}$$

whence:

$$\frac{R_{P.P.M.}}{R_{A.M.}} = -5 \text{ dB}$$

The more efficient method for demodulating P.P.M. described in Section 5.2 would, from Equation (7), result in:

$$\frac{R'_{P.P.M.}}{R_V} = +25 \text{ dB}$$

and $$\frac{R'_{P.P.M.}}{R_{A.M.}} = -2 \text{ dB}$$

The threshold value of pulse signal-to-noise ratio, below which the ratio $R_{P.P.M.}/R_{A.M.}$ falls catastrophically, occurs when the quasi peak-to-peak noise voltage is approximately equal to the pulse amplitude. For the system and pulse parameters assumed the threshold occurs at a picture signal-to-noise ratio of approximately 15 dB.

6. PRACTICAL CONSIDERATIONS

In this Section of the report the use of sound-modulated pulses combined with the video signal is discussed in terms of the two main applications mentioned in the Introduction, and some practical problems associated with each application are considered.

It is assumed, throughout the ensuing discussions, that the line-scan frequency of the television system is stable and free from phase and frequency modulation at audio frequency. Such a characteristic is desirable not only from the point of view of using sound-modulated pulses; it aids the design and performance of receivers and also of equipment having long mechanical time-constants such as that used by a broadcasting organization.

However, a broadcasting organization often requires to synchronize television signals obtained from widely separated sources and uses "genlock" and "slavelock" techniques for this purpose. These processes fundamentally involve phase and frequency modulation of the line-synchronizing signal

but, provided such modulation occurs at a suitably slow rate (which need not involve a significant increase in the time taken to perform the synchronizing process), the effect upon the pulse-sound system should be small. In the case of a properly designed slavelock system, the synchronization of a television-signal source takes place before the signal is used for programme purposes and the line-scan frequency is subject to only very small phase and frequency modulation after synchronization has occurred; with such a slavelock arrangement, the effect upon a pulse-sound system should be quite negligible.

6.1. 625-line Broadcasting

At the present time 625-line broadcasting in the U.K. is confined to u.h.f. using Standard I, in channels that are 8 MHz wide; Fig. 15(*a*) illustrates the channel arrangements used. In order to provide satisfactory v.h.f. coverage of the U.K. with two 625-line programmes it would be necessary to use appreciably narrower channels.

If the coverage obtained with 7 MHz channels were considered adequate, two alternative vision and sound signal arrangements could be used. These are:

(i) A vision signal with a main-sideband bandwidth of 5 MHz and a vestigial-sideband bandwidth of 0·75 MHz, together with a sound signal using a frequency-modulated carrier (i.e. transmissions according to Standards B or G).

(ii) A vision signal with a main-sideband band-

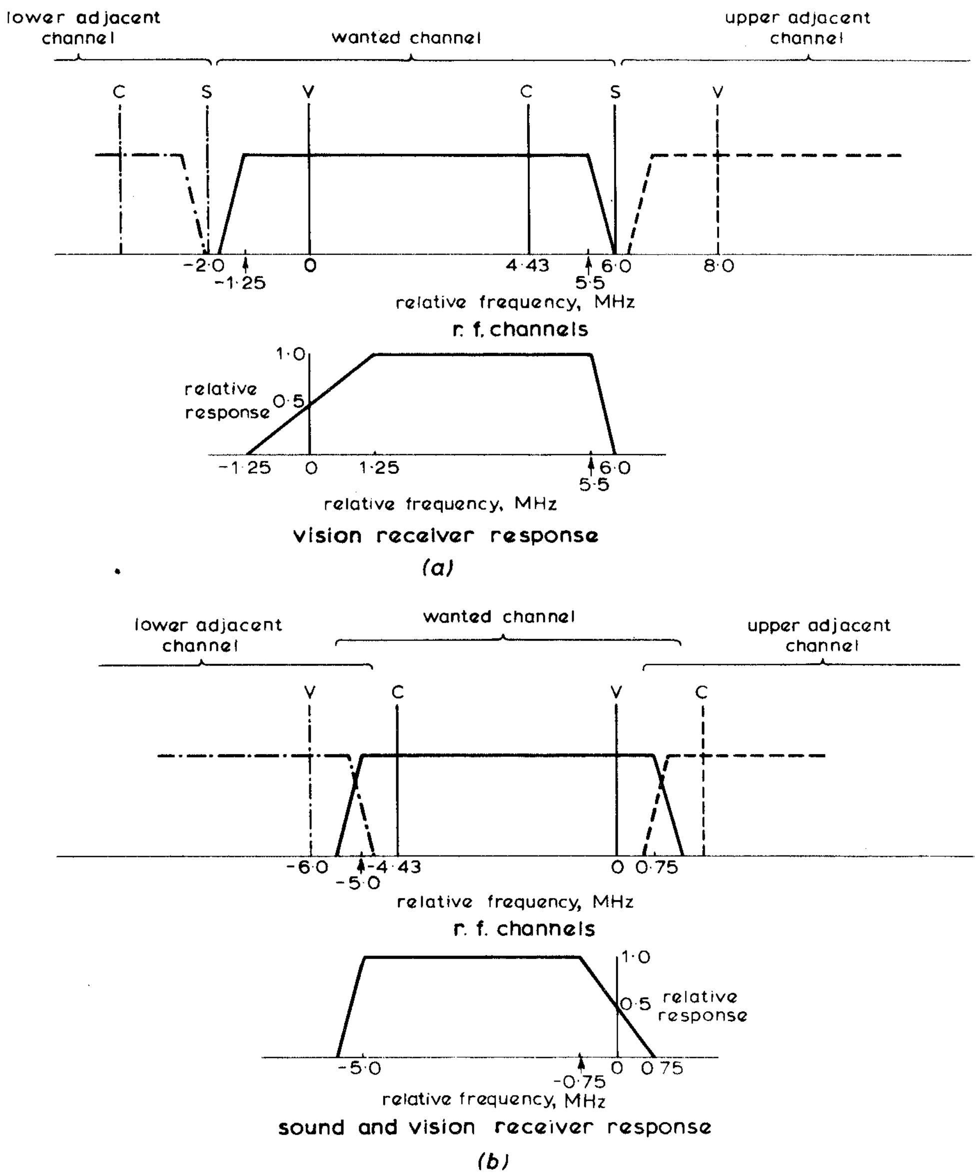

Fig. 15 - Current u.h.f. and possible v.h.f. channels
(*a*) R.F. channels and receiver response used in u.h.f.
(*b*) R.F. channels and receiver response possible, using pulse-sound, in v.h.f.
V = Vision Carrier, S = Sound Carrier, C = Colour Carrier

width of 5·5 MHz and a vestigial-sideband of 1·25 MHz, together with a sound signal using modulated pulses in the line-blanking intervals (i.e. vision transmissions according to Standard I with pulse sound).

Although Standards B and G are widely used in Europe, their narrower main and vestigial sideband bandwidths render the video performance of these standards inherently inferior to that of Standard I. Within the context of 7 MHz channel spacing, therefore, a vision signal conforming to Standard I with pulse sound offers some advantage.

If, however, the degree of coverage obtainable with 7 MHz channels were not considered adequate, it would be necessary to consider the use of still narrower channels with, say, 6 MHz bandwidth. Within such channels a vision signal similar to those of Standards B or G, together with pulse sound, is the only practicable arrangement that does not involve sacrificing the European subcarrier frequency of 4·43 MHz. This arrangement will be used as a basis for discussing the application of pulse sound to 625-line television broadcast transmissions. It should be emphasized, however, that the use of pulse sound is in no way responsible for the fact that the video performance provided by the vision signal is inferior to that provided by a signal conforming to Standard I.

Fig. 15(*b*) shows the relative disposition of adjacent channels. As will be seen, a certain degree of overlap between channels has been assumed; this appears practicable as the receiver is only required to reject the lower-adjacent vision carrier beyond the edge of the full (lower) sideband and the upper-adjacent chrominance signal beyond the edge of the vestigial (upper) sideband. Such a performance is not unduly different from that required by the arrangements of Fig. 15(*a*) where it is necessary to attenuate severely the sound signal of the wanted channel beyond the edge of the full (upper) sideband and to reject the lower-adjacent sound signal beyond the edge of the vestigial (lower) sideband.

In addition to a saving of channel bandwidth the elimination of the conventional f.m. sound transmission could confer other advantages. Present-day 625-line receivers employ intercarrier sound reception in which the beat between the sound and vision carriers is extracted from either the vision detector or the video amplifier and, as it is frequency modulated by the wanted sound signal, the beat is amplitude limited and demodulated in a conventional detector arrangement. Such a system demands that the sound carrier be passed, by the receiver i.f. amplifier, to the vision detector at a suitable level. If the level is attenuated by more than about 30 dB with respect to vision carrier, sound buzz occurs whenever the vision modulation contains strong 3 MHz components;[11] the sound buzz is due to the single-sideband demodulation by the vision detector of the 3 MHz vision-modulation components which, in turn, results in the production of 6 MHz harmonics that interfere severely with the sound intercarrier signal and may swamp it. Experience with 625-line monochrome transmissions shows that the design of present-day receivers is such that this effect can occur, although fairly infrequently, when the receiver is tuned correctly;[12] deliberate mistuning so as to increase the vision-carrier level at the vision detector with respect to the level of the 3 MHz sideband, thus inhibiting the production of 6 MHz harmonics, reduces the tendency to sound buzz by a large factor, at the cost of picture sharpness.

Further, the introduction of colour transmissions may bring to light a further difficulty. It has been shown[13] that, unless the sound-carrier input to the vision detector is attenuated sufficiently (i.e. > about 30 dB) with respect to the vision carrier, the beats between the sound carrier and the vision-signal sidebands corresponding to the chrominance signal (subcarrier frequency: 4·43 MHz) have an amplitude such as to impair the received picture significantly. These considerations assume that the receiver is correctly tuned and has the conventional asymmetric-sideband characteristic with a response at vision-carrier frequency -6 dB with respect to the peak response. Again, as in the case of sound buzz, the sound-to-chrominance beat may be considerably reduced in level (and thus rendered unobtrusive) by deliberately mistuning the receiver so as to decrease the level of the chrominance signal with respect to that of the sound signal; this may also result in a less sharp picture.

The above-mentioned effects may be avoided by suitable receiver design in which a separate detector is used to provide the sound intercarrier signal; the detector is fed from a point in the i.f. amplifier prior to the input circuit of the vision detector which may now be fitted with a sound-carrier trap.[12] Such an arrangement may well be used in colour receivers but has not yet found favour in the U.K. with designers of monochrome receivers, probably due to cost.

It may be deduced that, although the channe arrangements shown in Fig. 15(*a*) potentially provide a picture sharpness corresponding to a video bandwidth of 5·5 MHz, the effects of receiver mistuning to avoid either sound buzz or sound-to-chrominance beat (perhaps both!) may well result in a picture that is no sharper than that provided by the channel arrangements and receiver response shown in Fig. 15(*b*). Pulse-sound reception cannot suffer from the above-mentioned sound-buzz defect (sound and picture signals are not present simultaneously) and the beat that could occur between the wanted chrominance signal and the lower-adjacent vision carrier

is only likely under certain unusual reception conditions and, in addition, may be appreciably reduced in visibility by suitably offsetting the vision-carrier frequencies of adjacent channels; thus the tendency for the deliberate mistuning of receivers may be markedly reduced.

However, in considering the problems of 625-line broadcasting in the v.h.f. Bands, the main objective is to arrive at a plan whereby a single television-scanning standard may be used for all U.K. television broadcasting and it is important to consider how television transmissions incorporating sound pulses may be introduced without unduly complicating an already complex situation. Existing dual-standard receivers accept in v.h.f. 405-line transmissions, with positive vision-modulation and a.m. sound, and in u.h.f., 625-line transmissions with negative vision-modulation and f.m. sound.

One possible approach to the problem of introducing pulse sound is to consider its possible application only to future v.h.f. television broadcasting. In this case it would be necessary, prior to the introduction of 625-line broadcasting in v.h.f., to ensure that all receivers were capable of receiving:

(*a*) 405-line transmissions, with positive vision-modulation and a.m. sound, in v.h.f.

(*b*) 625-line transmissions, with negative vision-modulation and pulse sound, conforming to, say, 6 MHz channels in v.h.f.

and

(*c*) 625-line transmissions, with negative vision-modulation and f.m. sound, in u.h.f.

Thus it would be necessary to make available, to the public, receivers containing facilities which would not be used until some arbitrary future date when the 405-line transmissions in v.h.f. would be replaced by others using 625-lines; such receivers could well be appreciably more costly than existing dual-standard receivers and any failures in the circuits required only for v.h.f. 625-line transmissions would not be evident until the time of changeover, which could precipitate a rather desperate service situation. After the replacement of the 405-transmissions in v.h.f. by 625-line transmissions, receivers could then be marketed which were suitable only for the two types of 625-line transmissions (*b*) and (*c*) mentioned above. However, such receivers would be of dual-standard nature in that they would have to accommodate two different channel widths and two different types of sound modulation. It may be concluded, therefore, that such a method for changing over from 405- to 625-line broadcasting in v.h.f. would not receive enthusiastic support.

A second method of achieving a single television scanning standard appears feasible if modifications to the existing u.h.f. transmissions were considered practicable. This follows from the concept that, at some future date, it may be possible to use transmissions in both v.h.f. and u.h.f. which are identical, with the possible exception that 8 MHz channels are used in u.h.f. and, say, 6 MHz channels are used in v.h.f.; in such circumstances, the only form of sound transmission which could be used in both Bands is that employing pulses. In order to achieve this end, it would be necessary to introduce pulse-sound transmissions in u.h.f. in the near future, alongside the existing f.m. sound transmissions, and "transitional-stage" receivers could then be marketed which were capable of receiving:

(*a*) 405-line transmissions, with positive vision-modulation and a.m. sound, in v.h.f.

and

(*b*) 625-line transmissions, with negative vision-modulation and pulse sound, in both v.h.f. and u.h.f.

It would be necessary to ensure that these receivers ignore the f.m. sound transmissions radiated in u.h.f. for the benefit of existing dual-standard receivers. After the replacement of the 405-line transmissions in v.h.f. with 625-line transmissions, it would be possible to make available, to the public, receivers operating on only 625 lines with negative vision-modulation and pulse-sound in all Bands and, at this stage, the f.m. sound transmissions in the u.h.f. channels could be closed down. Further, once the u.h.f. f.m. sound transmissions were closed down it might prove possible, subject to considerations of international interference, to replan the u.h.f. Bands on the basis of, say, 6 MHz channels and thus provide scope for more programmes. These "final stage" receivers would be suitable only for the transmissions described in (*b*) above and although different channel widths might be used in v.h.f. and u.h.f., the absence of sound carriers would simplify design. If identical channel widths were used in all Bands after the close-down of the u.h.f. f.m. sound transmissions, this would enable truly single-standard receivers to be marketed; such receivers would probably prove appreciably cheaper than present-day receivers.

It will be appreciated that the practicability of utilizing this second method of achieving a single scanning standard is dependent upon it being possible to introduce pulse-sound transmissions into u.h.f. in the near future, without interfering with the operation of existing dual-standard receivers. This raises the issue of compatibility which, in turn, must influence the pulse parameters chosen for the pulse-sound system. The choice of a suitable form of pulse signal, bearing in mind

compatibility, and some of the factors in the design of pulse-sound demodulators for use in domestic receivers are dealt with in the following two subsections.

6.1.1. Form of Sound-Pulse Signal

Although Fig. 2 typifies one form of combined pulse-sound and vision signal, many factors must be borne in mind when specifying a broadcast signal which is to be introduced in the manner described in Section 6.1.

Principal requirements to be satisfied are:

(i) Existing 625-line receivers, when presented with the pulse-sound signal together with a conventional sound transmission, should provide pictures and sound that are not significantly different from those obtained at present.

(ii) The sound signal-to-noise ratio provided by pulse sound, when demodulated by a pulse-sound demodulator suitable for inclusion in a domestic receiver, must be satisfactory.

(iii) The sound quality, using a pulse-sound demodulator suitable for inclusion in a domestic receiver, must be satisfactory.

The first of the above-mentioned requirements influences, to a major extent, the location and shape of the pulse within the waveform. Any excursion of the pulse waveform below blanking level must not result in interference with receiver synchronization and, in order not to affect intercarrier sound reception, it is probably desirable to limit the excursion in the white direction at white level. Present-day 625-line receivers incorporate line-synchronizing circuits which may well be affected by additional pulses located within the line-synchronizing intervals, unless the fact that the pulses are sound modulated is undetectable at the output of the synchronizing-pulse separator. In such circumstances only P.A.M. could be used, with modulation that causes the pulse amplitude to vary between the limits of blanking level and white level; as may be deduced from Section 5.3.3, such an arrangement would have a poor noise performance. Further, few present-day 625-line receivers incorporate line-flyback suppression of the scanning beam in the picture display tube and, as a result, it is likely that sound pulses, located within the line-synchronizing intervals with peak excursions reaching white level, would cause visible impairment of the picture. These difficulties could be avoided if the sound pulses were located within the post-synchronizing line-blanking intervals, as suggested in at least one early proposal.[14] The resulting combined waveforms are illustrated in Figs. 16(*a*) and 16(*b*) where it will be seen that the field-synchronizing waveform has been modified in order to provide locations for the sound pulses similar to those occurring during line-blanking intervals. Figs. 16(*a*) and 16(*b*) assume that unidirectional pulses extending from blanking level to white level would be used and thus no interference with receiver line synchronization by sound pulses could occur; however, although the modified field-synchronizing waveform would provide, in principle, adequate field-synchronizing information, it would be necessary to ensure that present-day receivers would be unaffected. With regard to the visibility of the sound pulses on the screens of receivers, part of the post-synchronizing line-blanking interval coincides with the end of line-flyback in the receiver and it has been pointed out* that a sound pulse located at or near this point would occur when the scanning spot in the display tube of a properly adjusted receiver is behind the lefthand edge of the tube mask, and is thus hidden from the viewer. However, the early part of the post-synchronizing line-blanking interval has already been allocated to the colour burst and, in order to provide an adequate location for the sound pulse, it would probably be necessary to extend the interval somewhat,** thus reducing the duration of the active line; this would demand that the picture-width controls of existing 625-line receivers be adjusted slightly in order to place the sound pulse behind the tube mask.

With regard to the second requirement (i.e. a satisfactory signal-to-noise ratio) the results of Section 5.3 lead to the conclusion that P.T.M. should be used with the maximum possible pulse amplitude and deviation; problems of compatibility influence the choice between P.D.M. and P.P.M. In order to be able to use as high a pulse amplitude as possible, the pulse waveform should, if possible, traverse the region between blanking level and sync level; however, disturbance of receiver synchronization must be avoided. As may be deduced from Sections 3.2.1 and 3.2.2, the baseband-frequency components of a P.P.M. spectrum are of lower amplitude than those of the corresponding P.D.M. spectrum; as a result, a pulse train of the former type would be less likely to cause interference than one of the latter type. Further, a pulse occupying the preferred location whose waveform excursion reached white level and traversed the region between blanking and sync levels could be of such a shape as to give a spectrum with little energy at the low-frequency end of the video band. As it is customary, in domestic receivers, to restrict the bandwidth of the circuit connecting the video stage to the synchronizing-pulse separator, the high-energy components of the pulse would be severely attenuated prior to the separator, whose amplitude-limiting action would then prevent them affecting the receiver synchronization. Fig. 17 illustrates various forms of position-modulated pulse that could be used. In each case the pulse is bodily modulated in position;

* By J.R. Sanders.

** As suggested by F.C. McLean.

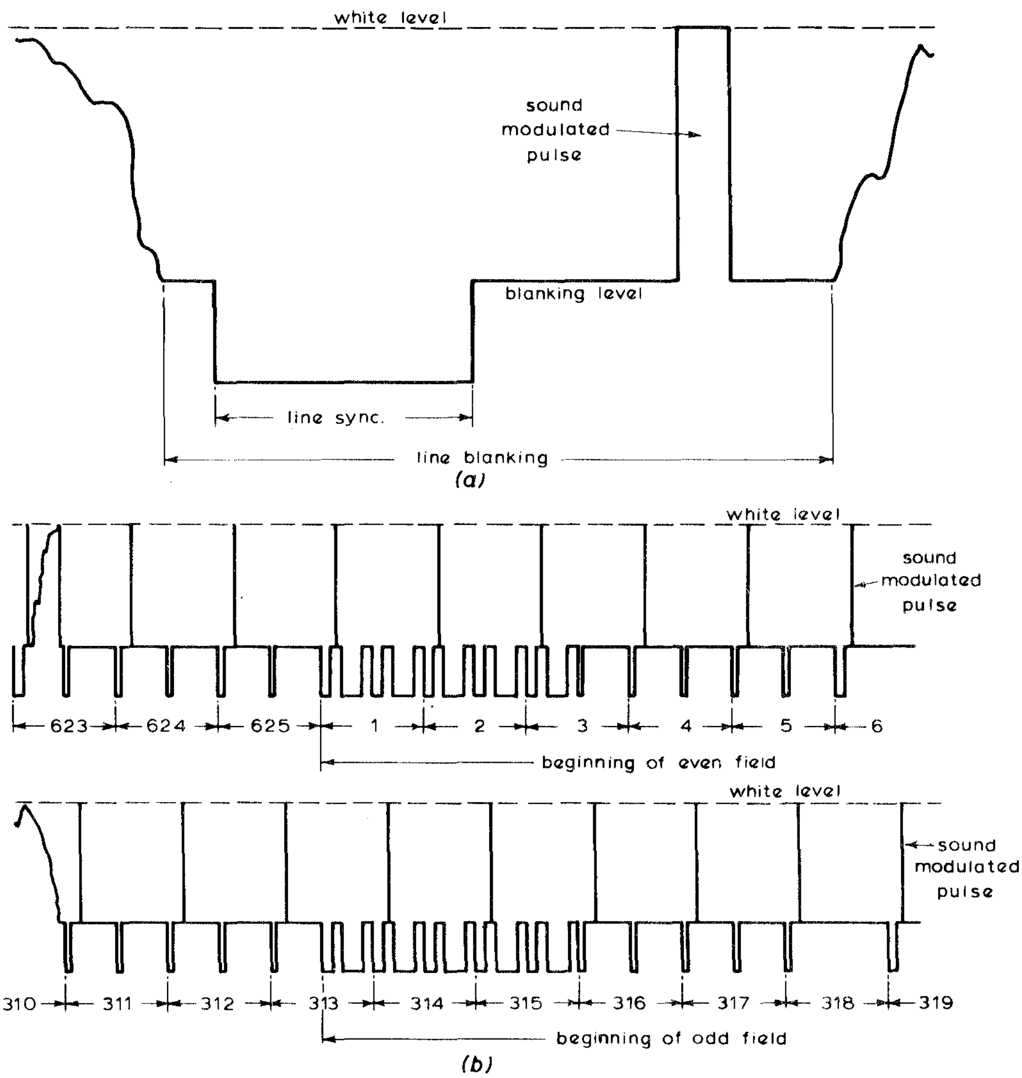

Fig. 16 - Sound-modulated pulses in post-sync line-blanking interval
(a) • waveform of line-blanking interval (b) waveforms of field-blanking intervals

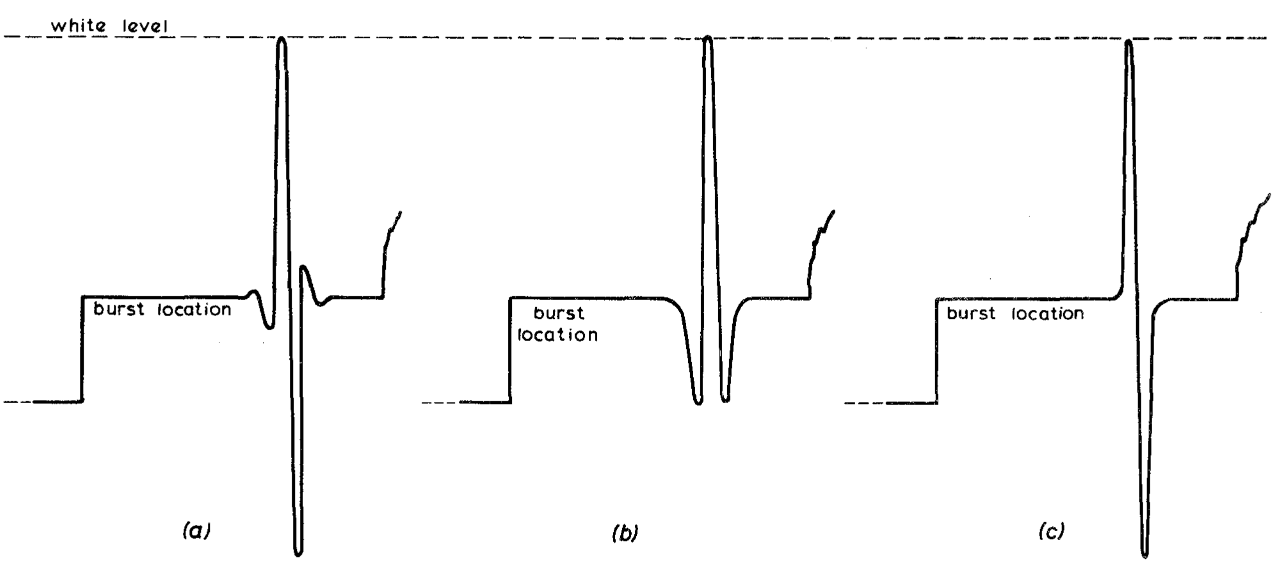

Fig. 17 - Some forms of position-modulated pulse suitable for broadcasting
(a) general form (b) "subtracted sine-squared" (c) "displaced sine-squared"

a "burst" formed from the modulation of a subcarrier by a position-modulated unidirectional pulse would not provide as good a signal-to-noise ratio. Waveform (*a*) is of a general nature which satisfies the requirements that it should have a large magnitude and be characterized by little energy at the low video frequencies. It will be seen that the pulse excursion extends well beyond sync level; this would probably be permissible using negative vision modulation. Waveform (*b*)* may be obtained by subtracting the waveform of a broad sine-squared pulse from that of a narrower one. By choosing suitable values of magnitude and duration for the two pulses it is possible to confine the overall waveform excursion within the limits of white level and synchronizing level whilst still forming a pulse with little energy at low video frequencies. Waveform (*c*)** is produced by combining two nominally identical sine-squared pulses; the second is delayed with respect to the first by approximately half the half-magnitude duration and is of opposite polarity.

The third requirement, that of satisfactory quality, must be considered in relation to the problems of providing demodulators suitable for inclusion in domestic receivers. Although, as described in Section 4.2.1, the use of uniform sampling could enable demodulators to be designed which gave intrinsically distortion-free sound within the band 0 to 7·8 kHz, it is likely that they would be appreciably more complex, and hence more expensive, than demodulators designed to operate with a signal based upon natural sampling. In view of the additional point that, with the relatively low value of pulse-position deviation which is possible in a combined vision and sound signal, the degree of distortion resulting from the use of natural sampling is probably quite tolerable, it would appear desirable to use natural sampling in a broadcast system. Further, as mentioned in Section 4.2.2, it appears practicable to devise pulse-modulation arrangements whereby the degree of distortion resulting from the use of relatively simple demodulators and natural sampling may be substantially reduced.

From the foregoing considerations it would appear that, for a pulse-sound signal that is most likely to prove both satisfactory and capable of introduction alongside the existing system, P.P.M. based upon natural sampling should be used. The pulse shape should be similar to one of those illustrated in Fig. 17 with values of duration, magnitude and deviation determined by many conflicting considerations, including compatibility, sound signal-to-noise ratio, sound-signal threshold conservation of the active-line period, and transmitter peak-modulation characteristics.

* Proposed by E.R. Rout.

** Proposed by J.R. Sanders.

6.1.2. Practical Demodulators

The basic principles of demodulation, for P.P.M. with natural sampling, have been described in Section 4.2.2. However, the pulse-sound section of a domestic receiver must perform several functions additional to actual demodulation.

As already mentioned, the combined sound and vision signal must be processed so as to derive a train of position-modulated pulses as free as possible from extraneous noise; thus, it would be necessary to gate the combined signal using a pulse, derived from the receiver line time-base, timed so as to provide a time-slot embracing the deviation range of the sound pulse. Further, in order to utilize the fast edge extending over the whole excursion of a pulse of the form shown in Fig. 17, it would be necessary to apply the gated pulse to a network (e.g. a differentiating network) whose output contains a major peak derived from the long fast edge; this peak could then be applied to a relaxation oscillator, as described in Section 5.2, in order to obtain a pulse suitable for application to the demodulator.

All three forms of demodulator described in Section 4.2.2 could, in principle, be used in domestic receivers. The arrangement consisting of a low-pass filter and correcting network is simple, and apparently cheap, but its sensitivity is such that considerable subsequent audio amplification would be required in order to provide adequate output. For the other two forms of demodulator it would be necessary to provide a train of pulses recurring regularly at the mean recurrence frequency of the sound pulses and suitably timed with respect to them. In addition, it would be highly desirable that the regular pulse train be as free as possible from perturbation by noise; otherwise the signal-to-noise ratio of the demodulator output would suffer. It might be expected that this regular pulse train could be obtained from the receiver line time-base. However some receivers utilize "hard-lock" time-bases whose sensitivity to noise would render them unsuitable while "flywheel" circuits tend to be perturbed, by field-synchronizing information, to an extent which would also render them unsuitable. Thus, it may be concluded that a pulse train derived from a typical receiver line time-base would be suitable only for providing time slots, as already discussed.

These difficulties could be overcome by deriving the regularly recurring pulses from the sound pulses themselves, using a suitable servo-controlled oscillator which is unresponsive to position modulation. Further, it would be possible to perform the process of ramp demodulation using the servo-control circuits. Fig. 18(*a*) shows an example of such an arrangement.[15]

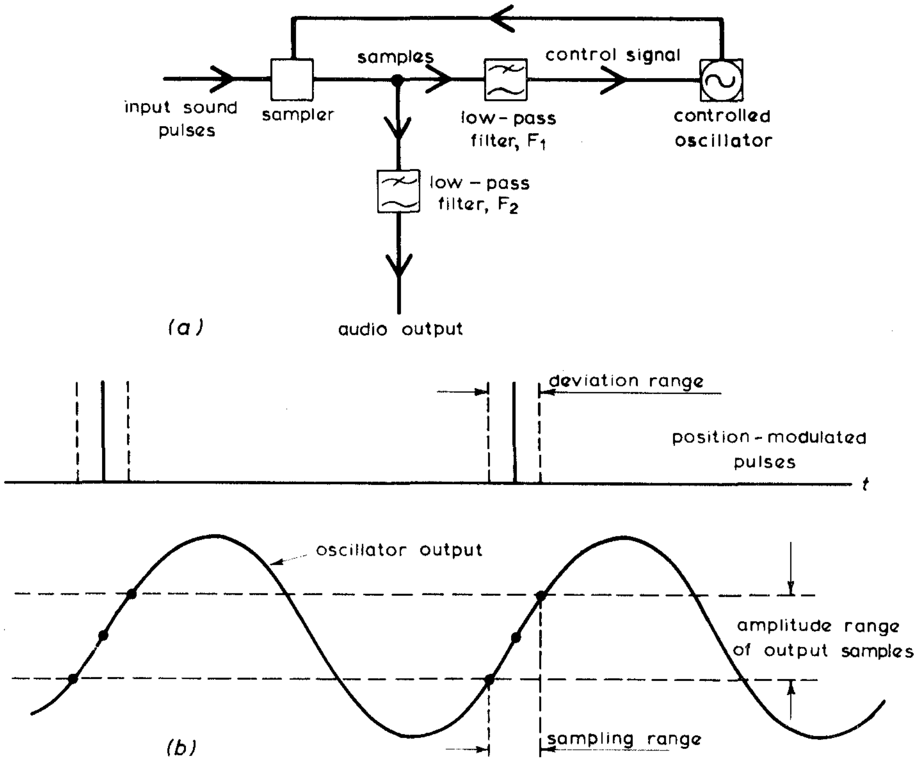

Fig. 18 - Ramp demodulation using a servo-controlled oscillator
(*a*) Block schematic (*b*) Sampling action and demodulation

Input sound pulses are used to sample the output from, say, a sine-wave oscillator whose frequency and phase are varied according to an applied control signal obtained by smoothing the output of the sampler in a low-pass filter F_1; such a servo-controlled oscillator is, of course, similar to a conventional flywheel time base. The oscillator operates at a frequency which is equal to (or is a low multiple of) the sound-pulse mean recurrence frequency and is sufficiently stable to ensure that, in the absence of input pulses, its frequency does not differ from the sound-pulse mean recurrence frequency (or a low multiple) by more than the cut-off frequency of the filter F_1. Thus, on applying unmodulated input pulses, correction of the oscillator frequency and phase takes place until the output of the sampler consists of identical samples of small (or zero) magnitude; this occurs when each input pulse samples corresponding points on high slope regions of the oscillator sine wave. However, if the input pulses are modulated, the phase of the oscillator varies in sympathy with the modulation unless the cut-off frequency of F_1 is made less than the lowest modulation frequency. By using such a low cut-off frequency the phase relationship between the oscillator output and the mean phase of the input pulses may be made constant. In such circumstances the sampler may now be regarded as part of a ramp demodulator and, by passing the samples to a low-pass filter F_2 having a cut-off frequency equal to half the mean pulse-recurrence frequency, the audio modulation may be recovered. Fig. 18(*b*) illustrates the operation of the arrangement for the case where the frequency of the oscillator is equal to the pulse mean recurrence frequency.

One further function should be incorporated in the pulse-sound channel of a domestic receiver. It is obvious that if the demodulator were presented with an erroneous input, such as picture signal, the output could cause loud and distressing noises from the loudspeaker; automatic muting circuits should be used to prevent this effect which may well occur whenever the receiver is grossly mistuned. Many methods of muting, of varying complexity and performance, may be devised, but simple and effective operation may well be obtained if the audio output of the demodulator were disconnected from the loudspeaker whenever its peak-to-peak magnitude substantially exceeded that available when the demodulator input is correct.

6.2. Sound Distribution using Vision Links

The ability to convey the associated sound signal along with the picture signal by means of time-division multiplex, using vision links which previously were used only for picture signals, would confer advantages to broadcasters and "common-carrier" authorities. The cost of separate sound circuits would be avoided and the frequency of occurrence of operational errors might be significantly reduced. Links in which the picture and

sound signals are now transmitted by frequency-division multiplex, such as those involving communication satellites, could be reduced in cost as a result of the reduction in the bandwidth necessary to convey both signals. Various systems of this nature are feasible, most of which require that the waveform of the video signal carried by the link be made non-standard so as to accommodate a pulse-sound signal giving a good sound-channel performance; in such circumstances, it would be necessary to modify the video waveform at the output of the link so as to conform with the standards used for broadcasting. The following two sub-sections outline possible forms of pulse-sound signal suitable for combination with link video signals and discuss some of the instrumentation involved.

6.2.1. Form of Sound Pulse Signal

A pulse-sound system for use with vision links should, if possible, satisfy the following requirements:

(i) The quality of the sound obtainable should be similar to that obtainable at present using separate sound circuits.

(ii) The signal-to-noise ratio at the output of the sound channel should be similar to that obtainable at present using separate sound circuits.

(iii) The presence of the sound signal should not increase the total signal excursion.

It will be inferred, from the first requirement, that the pulse-repetition frequency should be greater than the line-scan frequency of the 625-line 50-field system. Two pulses per line-blanking interval could be used, as mentioned in Section 2, thus permitting a maximum audio bandwidth of 15 kHz. As a very high degree of sound quality is desirable, the use of uniform sampling would appear appropriate.

The second and third requirements infer that the chosen form of pulse should provide as high a signal-to-noise ratio as possible. This consideration, together with the fact that the combined vision-and-sound signal might be subjected to non-linear amplitude distortion in the vision link, leads to the conclusion that P.T.M. should be used.

Two of the many possible forms of signal are illustrated in Fig. 19. It will be seen that, in both cases, P.P.M. has been assumed. Other possible forms of signal could employ P.D.M. For example, the use of asymmetrical P.D.M. with leading-edge modulation would ease the problem of deriving correct line-synchronizing information[3] at the output of the link;* on the other hand, the use of P.D.M. would substantially increase the amplitude of the audio-frequency component in the combined-signal spectrum which, in turn, could increase the possibility of interference with the picture signal in the presence of certain forms of distortion.

* In order to facilitate removal of the sound pulses and modification of the video waveform to the standard form.

The form shown in Fig. 19(*a*) employs two unidirectional pulses located in each line-synchronizing interval and in each corresponding interval within the field-blanking interval; such an arrangement may be used with a video signal whose waveform is conventional except that the number of equalizing pulses is reduced to one,[16] occurring during alternate field-blanking intervals. The signal-to-noise performance obtainable may be estimated by assuming that two sine-squared pulses are used, each with a half-amplitude duration of $0{\cdot}2\ \mu s$ and a peak deviation of $0{\cdot}5\ \mu s$.** Reference to Section 5.3.4 shows that, for such pulse parameters and double-edge demodulation, the r.m.s. audio signal-to-r.m.s.-noise ratio exceeds the picture signal-to-r.m.s.-noise ratio by 15 dB; although the audio bandwidth of the arrangement has been increased by a factor of two as compared with the example described in 5.3.4, this has been compensated by the increase in pulse recurrence frequency. As a typical good-quality vision link provides a picture signal-to-noise ratio of at least 46 dB, the r.m.s. audio signal-to-r.m.s.-noise ratio at the output would be 61 dB.

The second possible form of signal, illustrated in Fig. 19(*b*), utilizes one unidirectional pulse located in each line-synchronizing interval (and in each corresponding interval in the field-synchronizing interval) together with a further pulse in each post-synchronizing line-blanking interval which is characterized by one edge that extends from white level to synchronizing level and has a similar form to the pulse illustrated in Fig. 17(*c*), except for the excursion below blanking level.*** Thus, for such a signal, it would be necessary to use a video waveform having only one equalizing pulse per field-blanking interval and field-synchronizing pulses of the form shown in Fig. 16(*b*). The signal-to-noise performance may be assessed by assuming that the first (synchronizing-interval) pulse has a sine-squared shape, a half-amplitude duration of $0{\cdot}2\ \mu s$ and a peak deviation of $1{\cdot}5\ \mu s$ (as in the example of 5.3.4), and that the second pulse has a total duration of approximately $0{\cdot}6\ \mu s$, a long edge similar to the trailing edge of the first pulse and a peak deviation of $0{\cdot}5\ \mu s$.**** In such circumstances the r.m.s. audio signal-to-r.m.s.-noise ratio is mainly determined by the peak deviation of the second pulse. Owing to the fact that the peak deviation of the second pulse is only one-third that of the first pulse, the r.m.s. audio signal-to-r.m.s.-noise ratio is less than that which would have been obtained using the same peak deviation by a factor of 7 dB.

** Probably the maximum practicable assuming that no increase of line-synchronizing pulse duration is permissible.

*** Suggested by E.R. Rout.

**** Probably the maximum practicable assuming that a colour burst may be present and that no increase of line-blanking duration is permitted.

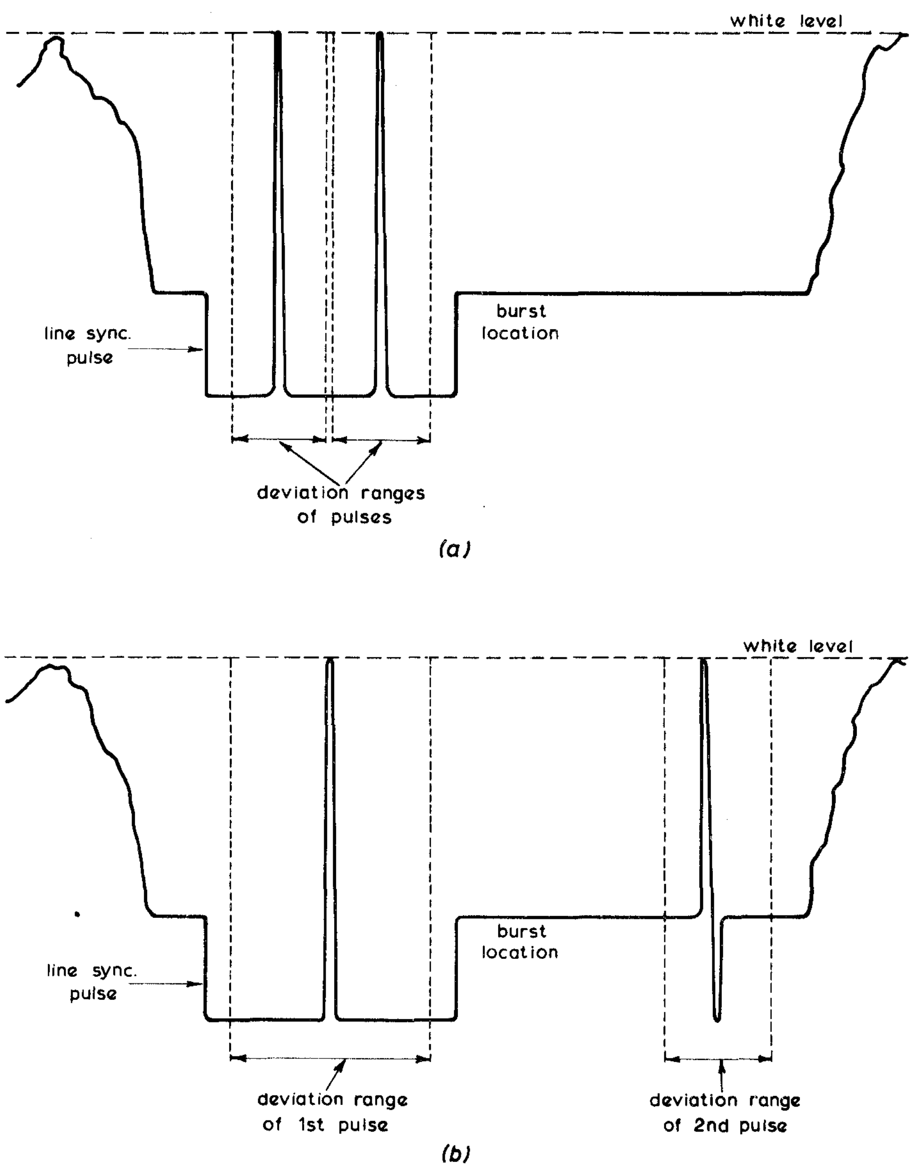

Fig. 19 - Two possible two-pulse arrangements
(*a*) Two-pulses per line-synchronizing interval
(*b*) One pulse per line-synchronizing interval plus one pulse per post-sync line-blanking interval

Thus, for double-edge demodulation the difference between the r.m.s. audio and picture signal-to-noise ratios would be approximately 18 dB. For a picture signal-to-r.m.s.-noise ratio of 46 dB the r.m.s. audio signal-to-r.m.s.-noise ratio would approximate to 64 dB. Thus, both forms of two-pulse signal could provide fairly similar performances in the presence of link noise. Both forms could provide somewhat better performances if the pulses used were characterized by higher edge slopes; this would probably be practicable for a link having a good performance with regard to bandwidth and freedom from group-delay distortion and might provide an improvement of 3 dB in each case.

Some further slight improvements could be obtained, if it were permissible to restrict the maximum audio frequency to, say, 10 kHz; in such circumstances the peak deviation of each of the two pulses used in the first form of signal could be increased by 1 dB without the time interval separating the two pulses ever falling below the previous value (i.e. for a maximum audio frequency 15 kHz). However, it must be borne in mind that a further reduction in the maximum audio frequency to, say, 7·5 kHz would enable a signal in the form described in Section 5.3.4 to be used, with a corresponding improvement in signal-to-noise performance of 10 dB. However, in assessing the significance of a change of signal-to-noise ratio due to a change of a.f. bandwidth, it must be borne in mind that the subjective effects of noise components having frequencies greater than 10 kHz are less than those with frequencies between 1 kHz and 10 kHz.

No consideration has been given to the problems, such as sound-system threshold, which could arise in the presence of high link noise; threshold would not occur until the picture signal-to-noise ratio fell well below a value at which the picture would be regarded as unsatisfactory.

6.2.2. Instrumentation

Most of the problems of instrumentation involved in a pulse-sound system for use with vision links are similar to those already described, in general terms, in Section 4. However, a system involving two pulses per line-blanking interval poses an additional problem not already mentioned; this Section, which outlines the generation and demodulation processes for sound signals of the type described in the preceding Section, describes the additional problem and poses a possible solution.

Generation of a signal such as that illustrated in, say, Fig. 19(*a*), may be performed as described in Section 4.2.1 and Fig. 6, using a sampling frequency equal to twice the line-scan frequency (approximately 31 kHz) and free from any phase modulation. The phase of the sampling frequency is adjusted so that alternate output pulses become the later of the two pulses shown in Fig. 19(*a*). The remaining pulses are then delayed by an interval almost equal to half a line period, thus providing the earlier pulse of each pair.

Immediately prior to demodulation it is necessary to delay the later pulse of each pair by an interval equal to the delay suffered, after generation, by the earlier pulses, thus reforming a conventional pulse train. Demodulation may then be carried out by a method such as that described in Section 4.2.1 and Fig. 7; the regularly recurring ramp waves, at twice the line-scan frequency, may be derived using the line-synchronizing pulses of the video waveform. However, in order to avoid difficulties due to possible slight phase modulation of the line-synchronizing pulses, a demodulator may be based upon the arrangement shown in Fig. 18 which utilizes the sound pulses themselves in order to provide the required regular ramp waves; as uniform sampling has been assumed, the output of the sampler must be applied to a "hold" circuit and resampled by a regular train of pulses, derived from a suitably delayed version of the oscillator output, before being applied to the low-pass filter.

The problem poses by such a system arises due to the need to delay the later pulse of each pair immediately prior to demodulation. If this delay does not equal that inserted in the earlier pulses immediately after generation, the output of the demodulator will include components at line-scan frequency together with unwanted inversions of the modulation. This could be avoided by applying the output from the demodulator, obtained from a point prior to the low-pass filter, to a sampler whose other input consists of regular pulses at line-scan frequency derived, say, from line-synchronizing pulses. The output of the sampler may then be smoothed in a circuit of long time-constant to give a control signal for adjusting the value of the delay. The smoothing ensures that 15 kHz components in the modulation do not actuate the delay control; complete avoidance of this difficulty could, of course, be obtained by restricting the upper frequency limit of the modulation to, say, 14 kHz.

After the sound signal has been extracted at the receiving-end of the link, it will normally be necessary to derive, from the combined signal, a video signal of the form suitable, say, for broadcasting. This process may be carried out using conventional techniques in which a local generator of standard synchronizing pulses is locked to the synchronizing pulses of the combined signal so that the line- and field-synchronizing pulses from both sources are respectively coincident (e.g. by "genlock"). In the case of a composite signal, as illustrated in Fig. 19(*a*), the original synchronizing pulses may then be removed and replaced by the standard form in a circuit which may be regarded as

a single-pole, double-throw, electronic switch; for the form of composite signal shown in Fig. 19(*b*), it would be necessary to replace both the synchronizing pulses and the blanking intervals.

7. CONCLUSIONS

A review of the properties of modulated pulses with particular reference to their application to television-sound transmission shows that suitable systems may be devised which would have considerable advantages over current arrangements.

For U.K. broadcasting, the use of a pulse-sound system employing one pulse per line-blanking interval would enable the current 405-line transmissions in Band I and III to be replaced by 625-line transmissions without drastically reducing the number of channels available. Further, it would appear feasible to introduce such a pulse-sound system alongside the conventional sound system used with 625-line transmissions thus enabling a smooth transition to be made from one sound system to the other without rendering useless any of the receivers in the hands of the public. For countries wishing to introduce bilingual transmissions, this form of pulse-sound system could provide the "second-language" channel without affecting the conventional sound channel.

The disadvantages of the pulse-sound system considered for broadcasting would be:

(i) The audio-frequency bandwidth would be restricted to about 7 kHz.

(ii) The audio signal-to-noise ratio under given conditions would be appreciably worse than that provided by the conventional f.m. system used at present with 625-line transmissions.

Neither of these disadvantages is considered serious. Very few present-day television receivers are capable of an audio output with a frequency range extending beyond 7 kHz and the signal-to-noise ratio provided by pulse-sound would be similar to that provided by the conventional a.m. system at present used with U.K. 405-line transmissions in the v.h.f. band, except under conditions where the picture signal-to-noise ratio was so low that the picture would be considered worthless.

The application of pulse-sound so as to enable the sound and vision signals to be transmitted, as a combined signal, through links appears quite feasible. The performance of the sound channel in such a system could be made comparable with that already provided using separate circuits and very significant economies could be made both by broadcasters and by common-carrier authorities. Operational advantages might also be obtained due to the ability to consider, at all times, the combined sound-and-vision signal as one programme signal which may be treated, in broad terms, as a video signal.

Experimental work, with particular reference to the application for broadcasting, is now well advanced and shows very promising results. Further work, directed to the link applications, appears to be desirable.

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APPENDIX I

The Signal-To-Noise Ratios of P.P.M. and Asymmetric P.D.M.

In this analysis it is assumed that each pulse of a P.P.M. train, after the removal of all noise located on all low-slope regions of the waveform, is used to initiate one edge of a pulse whose other edge is free from disturbance and repeats at the mean pulse-recurrence frequency, thus forming the corresponding P.D.M. train. The P.D.M. pulse train is then demodulated using a low-pass filter cutting off at half the mean pulse-recurrence frequency. This process is illustrated in Fig. 20(*a*) wherein it is assumed that the input P.P.M. pulses have unit magnitude and are accompanied by noise of r.m.s. value V_N (volts). Slicing and conversion to P.D.M. results in the waveform, again of unit magnitude, which is illustrated in Fig. 20(*b*).

The r.m.s. noise V_N is related to the mean power density $\overline{N}$, which is assumed uniform, and the video bandwidth f_0 by the relationship:

$$V_N^2 = \overline{N} \cdot f_0 \tag{10}$$

Further, the noise V_N may be considered to be the resultant of a very large number of sinusoids having equal amplitudes Δ_N, different frequencies ranging from zero to f_0, and random phase relationships. V_N is related to Δ_N by:

$$V_N^2 = \sum \frac{\Delta_N^2}{2}$$

Assuming that the noise voltage in a narrow frequency band df may be represented by a sinusoid of amplitude Δ_N, then, from Equation (10),

$$\Delta_N = \sqrt{2\overline{N} \cdot df} \tag{11}$$

From Fig. 20(*b*) it may be seen that the r.m.s. pulse-edge deviation ϵ, due to noise, is related to V_N by:

$$\epsilon = \frac{V_N}{S} \tag{12}$$

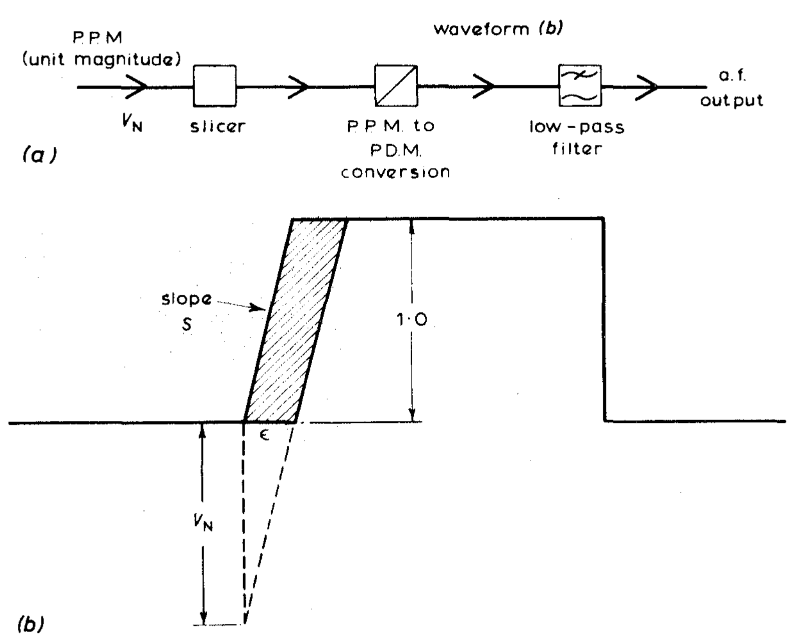

Fig. 20 - P.P.M. demodulator assumed for signal-to-noise calculation
(*a*) Slicing and demodulation : block schematic (*b*) Waveform of P.D.M. derived from P.P.M.

Hence the peak pulse-edge deviation due to one sinusoid of amplitude Δ_N is:

$$\Delta\epsilon = \frac{\Delta_N}{S}$$

From Equation (11) we have:

$$\Delta\epsilon = \frac{\sqrt{2\bar{N}\,.\,df}}{S} \qquad (13)$$

If the sinusoid of amplitude Δ_N has a frequency $\omega/2\pi$, then the spectrum of the pulse train, with a deviation of $\Delta\epsilon$ is, from Equation (4):

$$F'_N(\Delta\epsilon,t) = \frac{\Delta\epsilon}{T}\sin\omega_N t + \frac{1}{\pi}\sum_{m=1}^{\omega_0/\omega_r}\frac{1}{m}\left[\sum_{n=-\infty}^{\infty} J_n(\xi_m)\sin(m\omega_r + n\omega_N)t - \sin m\omega_r t\right]$$

where ξ_m is an abbreviation for $2\pi m\Delta\epsilon/T$

and ω_N may have any value between o and ω_0.

Assuming that ξ_m is very small compared with unity for all relevant values of m, we may take $J_0(\xi_m)$ as unity, $J_1(\xi_m) = -J_{-1}(\xi_m) = \xi_m/2$ and neglect entirely all Bessel functions of higher order.

$F'_N(\Delta\epsilon,t)$ thus simplifies to:

$$F'_N(\Delta\epsilon,t) = \frac{\Delta\epsilon}{T}\left[\sin\omega_N t + \sum_{m=1}^{\omega_0/\omega_r}\{\sin(m\omega_r + \omega_N)t - \sin(m\omega_r - \omega_N)t\}\right] \qquad (14)$$

The first term of Equation (14) has a uniform spectrum of amplitude $\Delta\epsilon/T$ at all frequencies between 0 and $\omega_0/2\pi$. The typical term $\sin(m\omega_r + \omega_N)t$ likewise has a uniform spectrum of amplitude $\Delta\epsilon/T$ at all frequencies between $m\omega_r/2\pi$ and $(m\omega_r + \omega_0)/2\pi$. The term $\sin(m\omega_r - \omega_N)t$ is a similar and balancing sideband equally spaced in the negative frequency sense, but it must be noted that $\omega_0 > m\omega_r$ except for the highest permissible value of m, so that for all values of $m < \omega_0/\omega_r$ the lower sideband suffers "folding" at zero frequency.

After demodulation by the low-pass filter cutting off at the frequency $f_r/2$(or $\omega_r/4\pi$) the total noise power at the output may be derived as follows:

The noise power contributed by the first term of Equation (14) is:

$$N_0^2 = \sum_{0}^{f_r/2} \tfrac{1}{2}\,.\left[\frac{\Delta\epsilon}{T}\right]^2$$

Substituting for $\Delta\epsilon$ from Equation (13) we have:

$$N_0^2 = \frac{\bar{N}}{S^2\,T^2}\int_0^{f_r/2} .\, df$$

$$= \frac{\bar{N}}{S^2\,T^2}\cdot\frac{f_r}{2} \qquad (15)$$

Owing to the "folding", the noise power contributed by each of the remaining terms (except the last) makes a double contribution, thus:

$$\sum_{m=1}^{\left(\frac{f_0}{f_r}-1\right)} N_1^2 = 2\left(\frac{f_0}{f_r} - 1\right)\frac{\bar{N}}{S^2\,T^2}\cdot\frac{f_r}{2} \qquad (16)$$

The noise power contributed by the last term is the same as that contributed by the first:

$$N_{f_0/f_r} = \frac{\bar{N}}{S^2\,T^2}\cdot\frac{f_r}{2} \qquad (17)$$

Whence the total noise power within the band zero to $f_r/2$ is

$$V_N'^2 = 2\frac{f_0}{f_r}\cdot\frac{\bar{N}}{S^2\,.\,T^2}\cdot\frac{f_r}{2}$$

$$= \frac{\bar{N}f_0}{S^2\,T^2}$$

$$= \frac{V_N^2}{S^2\,T^2}$$

Thus the r.m.s. noise output from the low-pass filter is:

$$V_N' = \frac{V_N}{S\,T}$$

The peak output signal corresponding to a peak deviation of δ (which corresponds to a peak deviation of d for P.P.M.) is, from Equation (4):

$$V_S = \frac{\delta}{T}$$

Whence the ratio of the r.m.s. signal to the r.m.s. noise is:

$$R_{\text{P.D.M.}} = \frac{\delta S}{V_N\sqrt{2}}$$

We may also write:

$$R_{\text{P.P.M.}} = \frac{d S}{V_N\sqrt{2}}$$

APPENDIX II

The Signal-To-Noise Ratio Obtained Using Intercarrier F.M. Reception, as Compared to That Obtained Using Only the Sound Carrier

If $1/k$ is the instantaneous ratio of the vision-carrier voltage to the sound-carrier voltage,*

v_{NV} is the instantaneous noise voltage accompanying the vision-carrier component at the intercarrier demodulator,

and V_{NS} is the r.m.s. noise voltage* accompanying the sound-carrier component at the intercarrier demodulator.

The relative effects of v_{NV} and V_{NS} upon the noise output of the demodulator may be expressed as:

$$v_{\text{NV}} = k \cdot V_{\text{NS}}$$

* Both the sound-carrier voltage and its accompanying r.m.s. noise voltage are assumed constant.

However, the factor $1/k$ varies as a function of the picture-signal magnitude. The mean noise power accompanying the vision-carrier component at the demodulator is:

$$\overline{v_{\text{NV}}^2} = \overline{k^2} \cdot V_{\text{NS}}^2$$

Hence the total mean noise power at the demodulator is:

$$V_{\text{NS}}^2 + \overline{v_{\text{NS}}^2} = V_{\text{NS}}^2(1 + \overline{k^2})$$

and the total r.m.s. noise input, V_{NT}, to the demodulator is:

$$V_{\text{NT}} = V_{\text{NS}}\sqrt{1 + \overline{k^2}}$$

$$= V_{\text{NS}}\sqrt{1 + K^2}$$

$$\text{whence } \frac{R'_{\text{F.M.}}}{R_{\text{F.M.}}} = \sqrt{1 + K^2}$$

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